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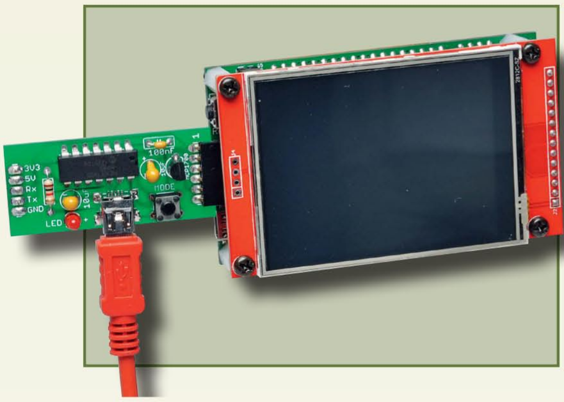

Projects and Circuits

HIGH PERFORMANCE RF PRESCALER 12
by Nicholas Vinen
If you have a frequency counter which will measure up to 10MHz, you can add this prescaler to provide a dramatic increase in performance – up to 6GHz.

MICROMITE BACKPACK V2 22
by Geoff Graham
This revised Micromite LCD BackPack incorporates the Microbridge, a USB interface and adds the ability to program the onboard PIC32 chip.

MICROBRIDGE 28
by Geoff Graham
The Microbridge is a tool for use with the Micromite, but it can also program any PIC32, or function as a USB-to-serial converter for the Arduino or Raspberry Pi.

Series and Features

TECHNO TALK by Mark Nelson 11
Bad broadband – Part 2

TEACH-IN 2018 – GET TESTING! – ELECTRONIC TEST EQUIPMENT AND MEASUREMENT TECHNIQUES 34
Part 8: Digital measurements

NET WORK by Alan Winstanley 46
Alexa – make me some money!... Maplin's sinking sands
A changing hobby... More smart talk

PIC n' MIX by Mike Hibbett 48
Practical DSP – Part 2

CIRCUIT SURGERY by Ian Bell 52
Source impedance and regulator stability

AUDIO OUT by Jake Rothman 57
Variable audio filters – Part 2: voltage-controlled filters

ELECTRONIC BUILDING BLOCKS by Julian Edgar 68
Digital Voltage Switch

Regulars and Services

SUBSCRIBE TO *EPE* and save money 4

***EPE* BACK ISSUES** 5

EDITORIAL 7
Jam tomorrow... Jam today?

NEWS – Barry Fox highlights technology's leading edge 8
Plus everyday news from the world of electronics

***EPE* TEACH-IN 8** 10

MICROCHIP READER OFFER 21
EPE Exclusive – Win a Microchip Curiosity HPC Development Board

***EPE* TEACH-IN 7** 33

***EPE* BACK ISSUES CD-ROM** 45

***EPE* PIC RESOURCES V2 CD-ROM** 56

***EPE* CD-ROMS FOR ELECTRONICS** 62
A wide range of CD-ROMs for hobbyists, students and engineers

DIRECT BOOK SERVICE 65
A wide range of technical books available by mail order, plus more CD-ROMs

***EPE* PCB SERVICE** 70
PCBs for *EPE* projects

ADVERTISERS INDEX 71

NEXT MONTH! – Highlights of next month's *EPE* 72

Teach-In 2018

Get testing! – electronic test equipment and measurement techniques

Our June 2018 issue will be published on Thursday 3 May 2018, see page 72 for details.

Readers' Services • Editorial and Advertisement Departments *7*

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PROJECTS • Micromite-Based Super Clock • Brownout Protector For Induction Motors • 100dB Stereo LED Audio Level/VU Meter – Part 2• **FEATURES** • Techno Talk • Teach-In 2017 – Introducing The BBC micro:bit – Part 2 • Interface • Net Work • PIC n' Mix • Circuit Surgery • Audio Out • Electronic Building Blocks • Max's Cool Beans

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DECEMBER '17

PROJECTS • Build Yourself A Digital Theremin • Precision Voltage And Current Reference With Touchscreen Control - Part 2 • Micromite Plus LCD Backpack • Micromite Plus Advanced Programming – Part 2 • **FEATURES** • Techno Talk • Teach-In 2018 – Get Testing! Electronic Test Equipment and Measurement Techniques – Part 3 • Net Work • PIC n' Mix • Circuit Surgery • Audio Out • Max's Cool Beans

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PROJECTS • Stationmaster • SC200 Amplifier Module – Part 3 • Using Cheap Asian Electronic Modules – Part 2 & 3 • **FEATURES** • Techno Talk • Teach-In 2018 – Get Testing! Electronic Test Equipment and Measurement Techniques – Part 6 • Net Work • PIC n' Mix • Circuit Surgery • Audio Out • Electronic Building Blocks •

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Jam tomorrow

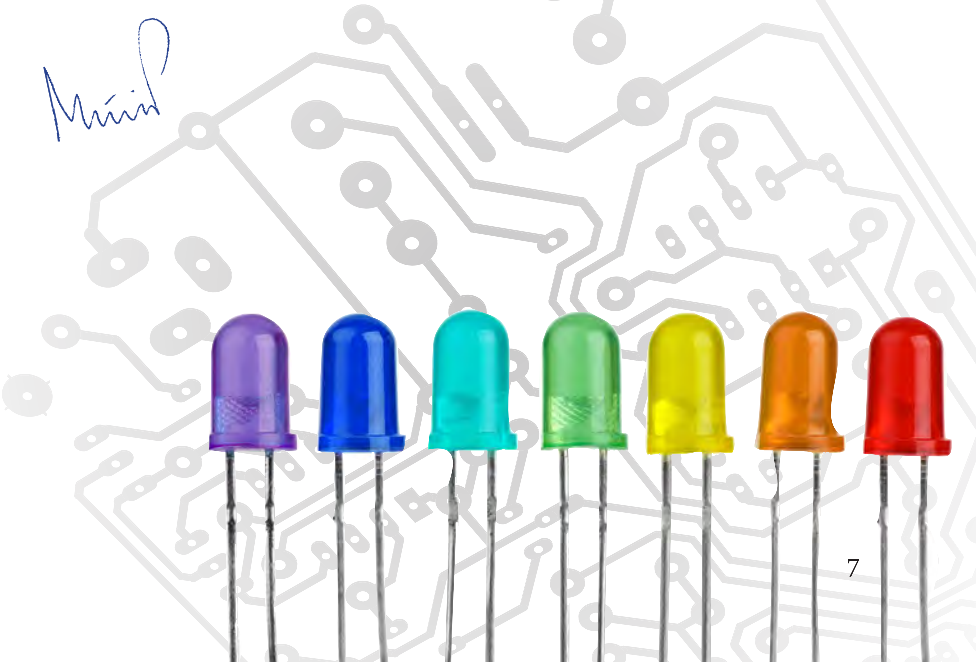
'Electricity too cheap to meter' was the optimistic prediction of nuclear energy proponents in the 1950s. Although this quote is often thought to refer specifically to nuclear fusion, it was really a reflection of the general expectation that science and technology would one way or another harness the power of the atom to supply near-limitless quantities of ultra-cheap electrical power.

The reality has tuned out rather different. Nuclear fission has undoubtedly produced a lot of kilowatt-hours, but even if we ignore the controversial aspects of this technology, it has certainly never been cheap. Nuclear fission is now a mature technology, but despite nearly three generations of engineering experience in building and running nuclear power stations, the electricity produced is expensive. Hinkley C, the huge reactor earmarked for the Somerset coast is expected to deliver an eighth of the country's electricity, but at £92.50 per megawatt-hour it is certainly a long way from 'too cheap to meter'. By comparison, the price of electricity from offshore windfarms has now plummeted to £57.50 per megawatt hour and is expected to fall still further. It is true that the wind does not always blow, but if the 'fuel' is free and the overall price is low, then it is surely a technology worth pursuing.

Jam today?

Can nuclear ever deliver on its 'too cheap' claim? Of course the real prize is not nuclear fission, but nuclear fusion. The dream of building a mini sun that essentially runs off seawater has been chased for decades, but the technical challenges of containing a highly controlled version of a hydrogen bomb are huge. The cynical critique of fusion power is that it's about 15-20 years from success – and always will be. It's easy to be cynical, but it doesn't solve problems. Time and again, what seemed impossibly difficult has been transformed by the unexpected appearance of a disruptive technology or technique.

Fusion researchers at MIT in the US are now working with, what they hope will be, a disruptive material. The fuel in fusion reactors operates at hundreds of millions of degrees and the only possible route to containing this plasma is in a 'magnetic bottle'. This bottle requires hugely powerful, current-thirsty magnets. MIT's approach is to work with magnets made from a newly available superconducting material – a steel tape coated with a compound called yttrium-barium-copper oxide (YBCO). They expect it to drastically reduces the cost, timeline, and complexity required to build a successful reactor. We'll see… in about 15 years.



NEWS

A roundup of the latest Everyday News from the world of electronics

BBC Natural History Unit repurposes military technology – report *by Barry Fox*

Try taking photos or videos of wild animals and you will usually be sorely disappointed with the results. The annual Awards ceremony of the International Moving Image Society, held recently in London at the beautifully restored Regent Street cinema, graphically explained why.

Natural History Unit

The BBC Natural History Unit (NHU) has been based in Bristol since 1957 and current head Julian Hector, along with innovation producer Colin Jackson, shared some of its secrets with IMIS members ahead of the presentation of a Wildlife Photography award to Gavin Thurston by Sir David Attenborough.

Mil-spec stabilisation and drones

'Because animals can't be "directed" like actors, we now shoot in 5K or better and pan and track within the frame area at post-production stage' Julian Hector explained. Very long focus telephoto shots from a helicopter are only now possible thanks to a military – but recently declassified – gyro-stabilised camera pod, which is mounted outside the cabin and remotely controlled from inside. Earlier attempts at getting camera stability by flying the crew in a hot-air balloon failed spectacularly because of a lack of steering control that left an NHU crew stranded in a jungle tree-top.

The NHU is also starting to use drones developed for military surveillance. HD/UHD cameras are now small enough to be mounted on animals, he said.

Freezing and swimming

Photographing penguins in Antarctica at -40°C involved a three-man crew living for 11 months in a shipping container, with 200TB of hard disc storage for the UHD video content. On the coldest days their LCD screens literally froze up and cables got brittle and snapped.

For underwater shooting the NHU now uses closed-circuit rebreathing suits developed to let military personnel stay submerged for over three hours instead of 30 minutes. 'The other advantage' said Hector, 'is that the sound of bubbles from normal scuba gear scares away fish – it is like going into a quiet jungle and shouting'.

Time-lapse photography developed for the building trade, to track the construction of bridges and skyscrapers, is now being used by the NHU to capture the very slow motion of starfish on the seabed, with a network of underwater lights turned on and off in sync with still cameras shooting very-high-resolution images which allow in-frame panning in post-production. A time-lapse sequence of Sahara dune movements was shot over two years. 'Unfortunately there are no mains sockets in the desert' says Hector. So the NHU uses using solar panels to trickle charge batteries.

Using infra-red to shoot bats in ultra slo-mo in pitch black caves needed nine banks of LED lamps; the camera crew needed gas masks because the fumes from bat droppings would have instantly knocked them unconscious.

Tech Holy Grails

The 'Holy Grail' is a fully solar-powered wildlife unit, says Hector. 'Night time is the right time for wildlife shooting' he says, citing another Holy Grail; a CMOS image sensor that can shoot motion video with just moonlight or starlight. Currently, the NHU uses recently declassified image intensifiers or thermal imaging cameras developed for night-time military operations. But the resolution is very poor.

Asked what he saw as the most significant development in wildlife filming so far, Hector immediately said 'colour – which wowed viewers and changed the game.'

Asked then about the possibility of shooting colour at night instead of the black and white captured by IR, image intensifier or thermal imagers, Colin Jackson explained: 'It's almost there, but I hope we don't do it because what we humans see at night is monochrome' 'I totally disagree' interjected Julian Hector 'it would let us tell the story from the animals' point of view'.

BBC Natural History Unit repurposes military technology – *continued*

A treat for historic cinema lovers – the restored Regent Street cinema

Afterwards I asked Sir David Attenborough for his view. ‘Monochrome I think’ he said. ‘It’s what we see and what animals see’.

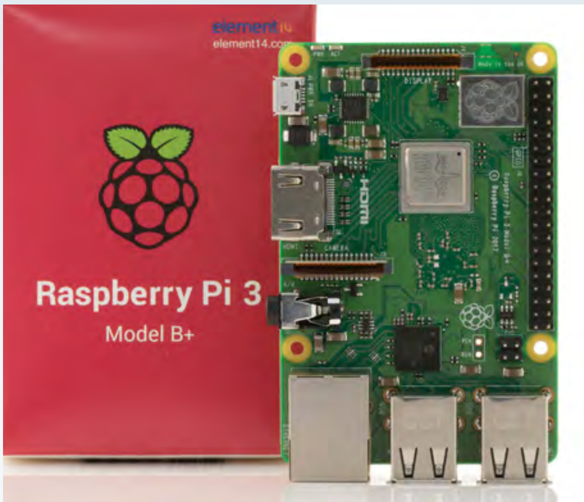
Cinématographe

For those with an interest in movie history, the Regent Street cinema is where in 1896 the Lumière brothers used their Cinématographe machine for the first movie presentation in the UK to a paying audience. The cinema closed to the public in 1980 and became a college lecture theatre, but was restored and re-opened as a public cinema by the University of Westminster in May 2015.

Power from Hammond Electronics

Hammond Electronics has extended its power distribution offering with an additional 12 variants of rack mounting and stand-alone 100-240VAC, 50/60Hz 10A power strips, designed for use with IEC power cords. For enhanced safety, two 10A resettable circuit breakers prevent overloading, and both types are available with either a double-pole single-throw green illuminated on/off switch or as a basic unswitched version with a green power-on indicator light. All are fitted with an IEC C14 inlet plug and multiple IEC C13 outlet sockets. Further details at: **www.hammfg.com/electronics/outlet-strips**

Launch of new Raspberry Pi 3 Model B+



Upgraded Raspberry Pi has a more powerful CPU, offers faster Ethernet and dual-band wireless networking

Farnell element14 has announced the launch of the Raspberry Pi 3 Model B+, the fastest and most powerful version yet, improving the already successful Raspberry Pi 3 Model B.

Built on a new quad-core Broadcom BCM2837 64-bit application processor running at 1.4GHz, the Raspberry Pi 3 Model B+ is over 15% faster than the Raspberry Pi 3 Model B.

Power over Ethernet (PoE) will be provided via a new official PoE add-on board for the Raspberry Pi, available from summer 2018.

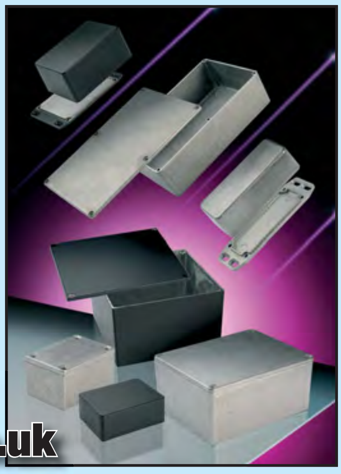
The Raspberry Pi 3 Model B+ is available as a standalone board and as an exclusive element14 Starter Pack, including a 16GB Micro-SD Card with NOOBS pre-installed, the official Raspberry Pi 2.5A power supply, and the official Raspberry Pi Case.

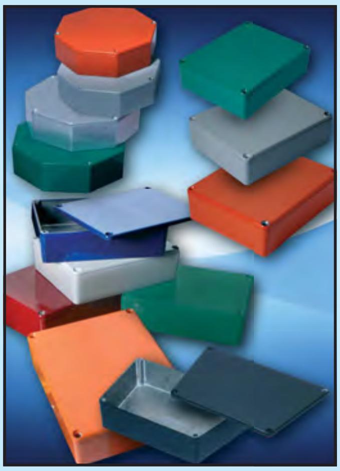
Pi 3 Model B+ offers backwards compatibility by following the same mechanical footprint as the Pi 2 and Pi 3 Model B, and includes:

- BCM2837B0, quad-core ARM Cortex-A53 64-bit SoC (1.4GHz)
- 1GB LPDDR2 SDRAM
- 2.4GHz and 5GHz IEEE 802.11ac wireless, Bluetooth 4.2, BLE
- Gigabit Ethernet over USB 2.0 (max throughput of 300Mbps)
- 40pin extended GPIO
- CSI camera port for connecting the Raspberry Pi camera
- DSI display port for connecting touch screen display
- 4 × USB 2 ports
- 4-pole stereo output and composite video port
- Full-size HDMI
- H.264, MPEG-4 decode (1080p30). H.264 encode (1080p30). OpenGL ES 1.1, 2.0 graphics
- Micro SD port for loading the operating system and storing data

The official PoE add-on board for Raspberry Pi 3 Model B+ includes:

- 802.3af PoE
- Class 2 device
- Fully isolated SMPS
- 36V–56V input voltage
- 5V output voltage, 2.5A output power
- Fan control





Bad broadband Part 2

TechnoTalk

Mark Nelson

Last month, we examined ways of improving sluggish broadband performance. A reader who prefers not to be name-checked wrote in to offer further helpful hints. Given that most of us value reliable and consistent broadband service for our hobby (and other) pursuits, I thought it would be useful to pass on this information, which comes 'straight from the horse's mouth' – from a time-served telecoms man.

IN OUR FRIEND'S EXPERIENCE, poor broadband speed is frequently not a network problem but the result of shortcomings in customers' pre-existing internal wiring, which was fine for speech but definitely sub-optimal for data transmission. Very often this is due to the presence of the 'blessed' bell wire (see: **www.filesaveas.com/jarviser/btiplate.html** to identify the wire in question).

More is less

Ever since Antonio Meucci invented the telephone (and Graham Bell took the credit, see **https://en.wikipedia.org/wiki/Antonio_Meucci**), telephones have used two wires (or one wire and earth) to carry speech and ringing. It was only when British Telecom introduced plug-in telephones in the 1980s that an extra wire was added to premises' wiring, to enable speech to be separate from ringing. It made sense then, but this so-called bell wire has no place or function in modern life. It is a throwback to the days of pulse dialling, induction coil telephones and magneto bells.

So, the advice is ditch this wire! Disconnect Terminal 3 on the removable faceplate of the NTE (master socket), and disconnect Terminal 3 connections at all extension sockets. Modern telephones are designed for two-wire operation, even BT-supplied ones. But why does the third (bell) wire cause problems? In one word, 'unbalance'.

Common (mode) problem

As we all know, common-mode noise (interference) can be present on all lines of your internal wiring. To clarify, common-mode means that it appears equally on both signal leads of a transmission line. If it is induced equally in both wires, there is no potential difference at the termination, meaning effectively no noise is detectable! If, however, we upset the carefully-created balance of an internal wiring cable (assured by its manufacturing design) by adding another wire to the circuit, the result is the telephone line becomes a radio antenna – because where we previously had just two wires, we now have three that are decidedly not equally balanced.

Long-wave broadcast signals and/or any other radio frequencies (from power transformers, PC monitor screens, fluorescent lights and suchlike) will all now be able to intermingle with the broadband signal. The broadband signal that comes down the telephone wiring at this point is still analogue, not digital as many suppose.

Says our friend, 'I have proved this point on faults. Measure the DSL line speed (not ISP speed) before work starts. To clarify, line speed (line rate) is the physical or 'sync' speed at which you router communicates with the equipment directly connected to the other end of the telephone line in the exchange. This is not necessarily what a speed tester website will indicate on your PC screen, although some speed testers show both.

'Now disconnect the wire at all Terminal 3 connection points. Then measure the line speed again. Presto! – a marked increase. The ISP speed will also rise, in time. In time, because the ADSL line is adaptive (based on a gradual learning process at the telephone exchange), the adjustment will take time, possible a day or more. The Line speed, or sync speed, responds immediately (in your hub manager).'

Are we nearly there now?

Hopefully, but not necessarily. Once airborne interference is eliminated from your internal 'premises' wiring, you have done all you possibly can. But strong interference on the external wiring can still drag down your broadband speed. It's sneaky stuff and detecting it is not an intuitive process.

Fortunately, there is any easy tool to help locate the source of the interference: an ordinary AM transistor radio. This will almost certainly be fitted with a ferrite rod aerial, which is directional. So switch on the tranny (indoors) and tune the medium wave to any quiet frequency on which you cannot hear a broadcast station. Quiet 'white' noise should be heard. If you do hear raspy, repetitive interference, you're onto something. This covers a broad spectrum, so the actual frequency you tune to doesn't matter. Once you get near the source of the trouble, the sound will become quite raucous.

Now you should rotate the radio slowly about a vertical axis. As **http://topbanddf.org.uk/whatis.htm** explains in greater detail, there will be two directions where the signal fades away. These nulls may be very narrow, but when located in the null, the aerial in your radio is pointing along a line joining your location and the location of the transmitter. The aerial is usually along the long axis of the radio when it is standing in its normal way.

Next, go outdoors and repeat this process on all four sides of your premises, until you have a good idea where the QRM (interference) is coming from. Timing might be critical. A machine giving off radio-frequency interference (RFI) may operate only at certain times, eg, during working hours, so a bit of detective work is sometimes needed.

'In one case we had,' says our informant, 'several customers in a particular rural location who were being affected on their broadband. The customers were dispersed, but they were all served by the same DP (telephone wire distribution pole). So, I walked the route with a transistor radio. Then, where the BT cables passed near a pole-mounted power transformer, the noise on the radio was intense. The transformer was from medium-voltage 11kV to low-voltage 230V. The electricity supply people were notified and they co-operated. Lo and behold, the noise ceased as soon as they removed the input links. They said the transformer was arcing internally.'

Not all broadband botherations will be as complicated or involved as this, but nearly all are soluble. If your broadband is driving you to desperation, why not have a word with your neighbours? If their troubles are as bad as yours, then you can press your broadband supplier(s) much harder.

Measure frequencies up to 6GHz and higher...

High Performance RF PRESCALER

by NICHOLAS VINEN

Would you like to measure frequencies up to 6GHz or more... but your frequency counter is not in the race? Well, if you already have a frequency counter which will measure up to 10MHz or so, you can add this prescaler to provide a *dramatic* increase in performance. And it has selectable frequency division ratios of 1000:1, 200:1, 100:1 or 10:1 to make it especially versatile.

A **frequency counter** is a very handy tool, even if it's one that's just built into a digital multimeter (DMM). Some DMMs contain frequency counters that will work up to 10MHz or more.

If you have one of those, or any other frequency counter (perhaps you built our low-cost 50MHz frequency meter from the November 2008 issue) – you can now have the facility to measure frequencies far above that range.

After all, there are lots of devices these days that operate at high frequencies – for example 433MHz, 900MHz, 2.4GHz or even 5.6GHz – so it's quite likely that you will soon want to measure the frequency of a signal and your cheap counter just won't be able to handle it. But now you can combine your existing frequency meter with our new *RF Prescaler* and you can get up into the gigahertz range.

The new *RF Prescaler* is housed in a tiny diecast aluminium case with two BNC output sockets and one SMA input socket. It also has a tiny 4-position slide switch to select the division ratio of 1000:1, 200:1, 100:1 or 10:1.

Set it to 1000:1 and connect it between the signal source and your meter and the 2.4GHz signal becomes 2.4MHz; easy for your meter to read and easy to convert in your head, since you just need to swap the units.

Operating principle

The basic arrangement of the *RF Prescaler* is shown in the block diagram of Fig.1 opposite. The source signal is applied to the 50Ω input connector at left, and then AC-coupled to IC1. This monolithic amplifier IC is essentially just a high-frequency Darlington transistor with biasing resistors and its input and output are both matched to 50Ω. 3.4V DC is fed to its collector via an inductor.

The output signal from the collector of IC1 is then AC-coupled to IC2, an identical amplifier, giving 22-34dB of signal boost in total, depending on frequency. The two amplifier stages are included to help make up for any signal loss in the input cabling and to give the *RF Prescaler* good sensitivity.

The output from IC2 is then fed to one of the differential inputs of a high-performance divide-by-five counter, IC3. The other differential input of IC3 is AC-coupled to ground since we don't actually have a differential signal at this point. IC3 is the most critical part of this circuit as it must reduce the very high frequency input signal down to something more manageable, ie, it gives a 1.2GHz output for a 6GHz input.

The output of IC3 is AC-coupled to another counter IC (IC4). This is programmable and can divide the frequency by a value anywhere between two and 256. Four different ratios are available, selected by slide switch S1: 2, 20, 40 and 200. These give overall division ratios (including the divide-by-five action of IC3) of 10, 100, 200 or 1000.

The output of IC4 is also differential, so these signals are fed to the bases of two PNP transistors which form a long-tailed pair. Their emitters are connected to the two output BNC connectors via impedance-matching resistor networks, which give an output impedance of 75Ω. Either or both outputs can then be fed to a frequency counter with a 50Ω or 75Ω input impedance. Or you could use one output to drive a frequency counter while the other drives an oscilloscope.

To handle the high frequencies involved, IC4 is an ECL (emitter-coupled logic) device with a maximum recommended operating frequency of 1.2GHz, although it will typically work up to 1.4GHz. IC1, IC2 and IC3 must all handle the full input frequency; so

they use heterojunction bipolar transistors (HBTs) to achieve operation up to around 8GHz.

IC1 and IC2 are made from indium gallium phosphide (InGaP) semiconductor material, rather than silicon, because electrons move through it more quickly. IC3 also uses InGaP, together with gallium arsenide (GaAs) semiconducting material.

The use of different semiconductor materials for the emitter-base and base-collector junctions allows the base to be much more heavily doped without creating excessive hole injection from the base to emitter. The heavier doping reduces the base resistance while maintaining gain. This is what the term ‘heterojunction’ refers to; ie, the fact that the transistor junctions are made from two *different* types of semiconductor.

The operation of the circuit is shown in the scope grab labelled Fig.2. The *RF Prescaler* has been set to its minimum 10:1 overall division ratio to better illustrate its operation. A 20MHz, 35mV RMS signal was applied to the unit and the output of amplifier stage IC2 is shown at the bottom of the screen in blue, with an amplitude of a little over 1V RMS. Overall gain is therefore 29dB [$20\log_{10}(1000 \div 35)$], within the range expected.

The output of divide-by-five prescaler IC3 is shown just above it in pink, and this is a fairly clean 4MHz square wave with an amplitude of about 500mV peak-to-peak. The signal from output connectors CON2 and CON3 are shown in green and yellow above, with the expected frequency of 2MHz and a peak-to-peak voltage of around 300mV.

Setting a division ratio of 100:1, 200:1 or 1000:1, the duty cycle of the outputs drops below 50%. The output pulse width is normally five times the input signal period, ie, with a 5GHz input, the output pulses are at least 1ns.

Fig.3 shows the unit operating with a 1000:1 division ratio and a 100MHz, 10mV RMS input signal. The mauve trace shows the output of amplifier IC2, with an RMS amplitude of 300mV, indicating a gain of around 29.5dB. As you can see, the output pulses are around 50ns and the output frequency is 99.99kHz, indicating that the input is actually just a little below 100MHz (ie, around 99.99MHz).

Features and Specifications

Input frequency range......5MHz-6GHz; typical operation to 7GHz
Input ....SMA, 50Ω
Input sensitivity ....<12mV RMS 6-3500MHz; <130mV RMS 5MHz-7GHz (typical; see Fig.6)
Division ratio ....selectable; 1000:1, 200:1, 100:1 or 10:1
Outputs ....2 x BNC, 50/75Ω, 180° out of phase
Output amplitude ....typically 300mV peak-to-peak into 50Ω
Output duty cycle ....approximately 50% (10:1), 5% (100:1), 2.5% (200:1), 0.5% (1000:1)
Output overshoot ....<10%
Power supply ....9V DC/500mA plugpack or 5V DC/500mA (microUSB); typically 375-450mA, quiescent ~375mA

Circuit description

The complete circuit of the 1000:1 *RF Prescaler* is shown in Fig.4. Input SMA connector CON1 is shown at left; depending on the exact model used, this can handle signals up to 20GHz. Low-capacitance schottky diodes D1 and D2 clamp the signal amplitude to no more than a few hundred millivolts to protect the rest of the circuit from a signal with too much amplitude. The signal is then AC-coupled via a 10nF C0G capacitor to the first amplifier, IC1.

IC1 is an ERA-2SM+ which provides around 16dB of gain at 1GHz, falling to 10.7dB at 6GHz. Its input impedance is 50Ω, so no termination resistors are required.

DC power is fed in via RF inductor L1, an ADCH80-A+, which maintains significant inductance up to 10GHz. It isolates the DC power supply from the AC signal present at output pin 3.

The 10nF bypass capacitor connected immediately adjacent to L1 helps to prevent any residual RF signal which may be coupled across L1’s small inter-winding capacitance from passing into the DC power supply.

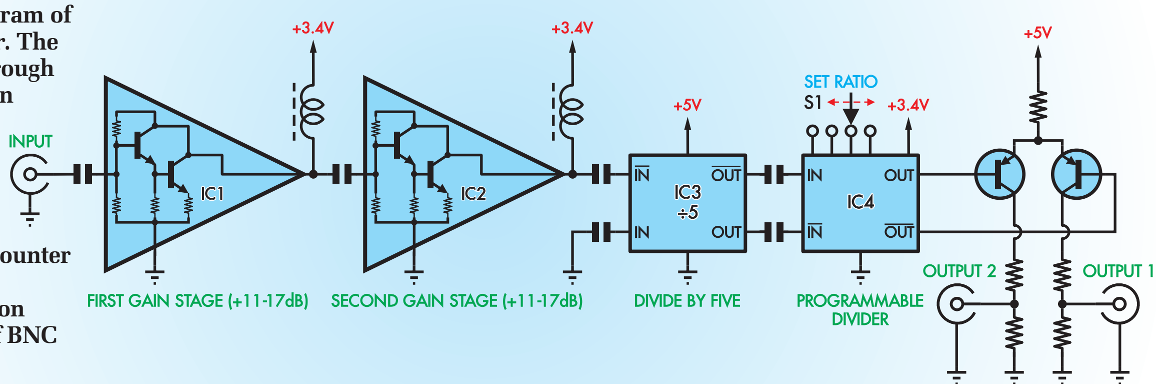
As the output impedance of IC1 is also 50Ω, we can feed its output signal directly to IC2 via another 10nF capacitor. The amplification stages comprising IC1 and L1 and IC2 and L2 are identical. Both amplifiers have a snubber network at their output comprising 33Ω resistors and 100pF capacitors. These help prevent instability when operating at around 4-4.5GHz.

The output from IC2 is fed to pin 3 of IC3 via another 10nF AC-coupling capacitor. IC3 is the HMC438MS8GE RF divide-by-5 counter and its differential input pins 2 and 3 are each internally biased and matched to 50Ω. As mentioned earlier, the other input terminal at pin 2 is connected to ground via an identical 10nF capacitor. Thus, this pin will sit at a DC level determined by IC3’s internal biasing network.

IC3 runs from a 5V supply which is smoothed by a low-pass filter comprising a 47μH inductor and parallel 10μF and 10nF capacitors. The 10μF capacitor provides bulk bypassing while the 10nF C0G capacitor has a much lower effective series inductance (ESL) and thus will be more effective at filtering out higher frequencies.

This filter helps prevent any high-frequency signals which may be present in the 5V power supply from upsetting the operation of IC3, and also prevents any modulation of its own supply current from being fed back into other components.

Fig.1: block diagram of the *RF Prescaler*. The signal passes through two amplification stages, then a differential divide-by-five prescaler, followed by a programmable counter and then a dual voltage conversion stage to a pair of BNC outputs.



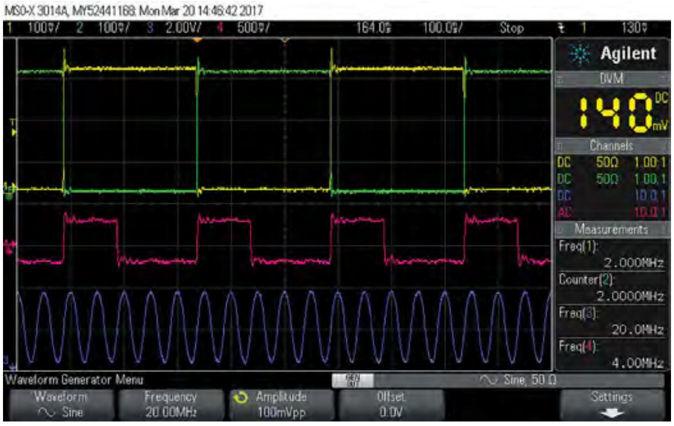


Fig.2: the amplified 20MHz input signal is shown at bottom in blue, followed by the 1/5 (4MHz) frequency signal above in pink and the 1/10 (2MHz) output signals at top, in yellow and green.

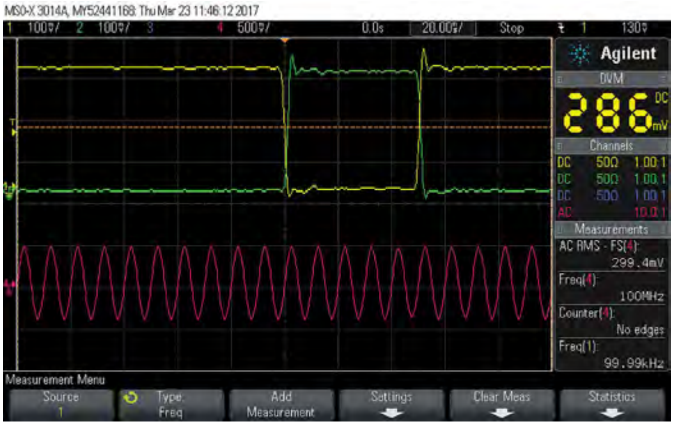


Fig.3: the pink trace shows the output of amplifier IC2 when fed with a 100MHz sinewave, and at top, the two outputs at 1/1000 the frequency, ie, 100kHz. The output pulses are around 50ns long.

IC3 can operate from very low frequencies (practically DC) up to around 7.5GHz, as shown in Fig.5. The upper limits shown here are not an issue since the 'saturated output power' of IC2, which provides the input signal for IC3, is 14dBm at 100MHz, 13dBm at 2GHz and 12dBm at 4GHz.

Hence, IC2 is incapable of producing a signal with an amplitude above that which IC3 can handle; we don't have data above 5GHz, but it seems probable that its output power is no more than 10dBm above this frequency.

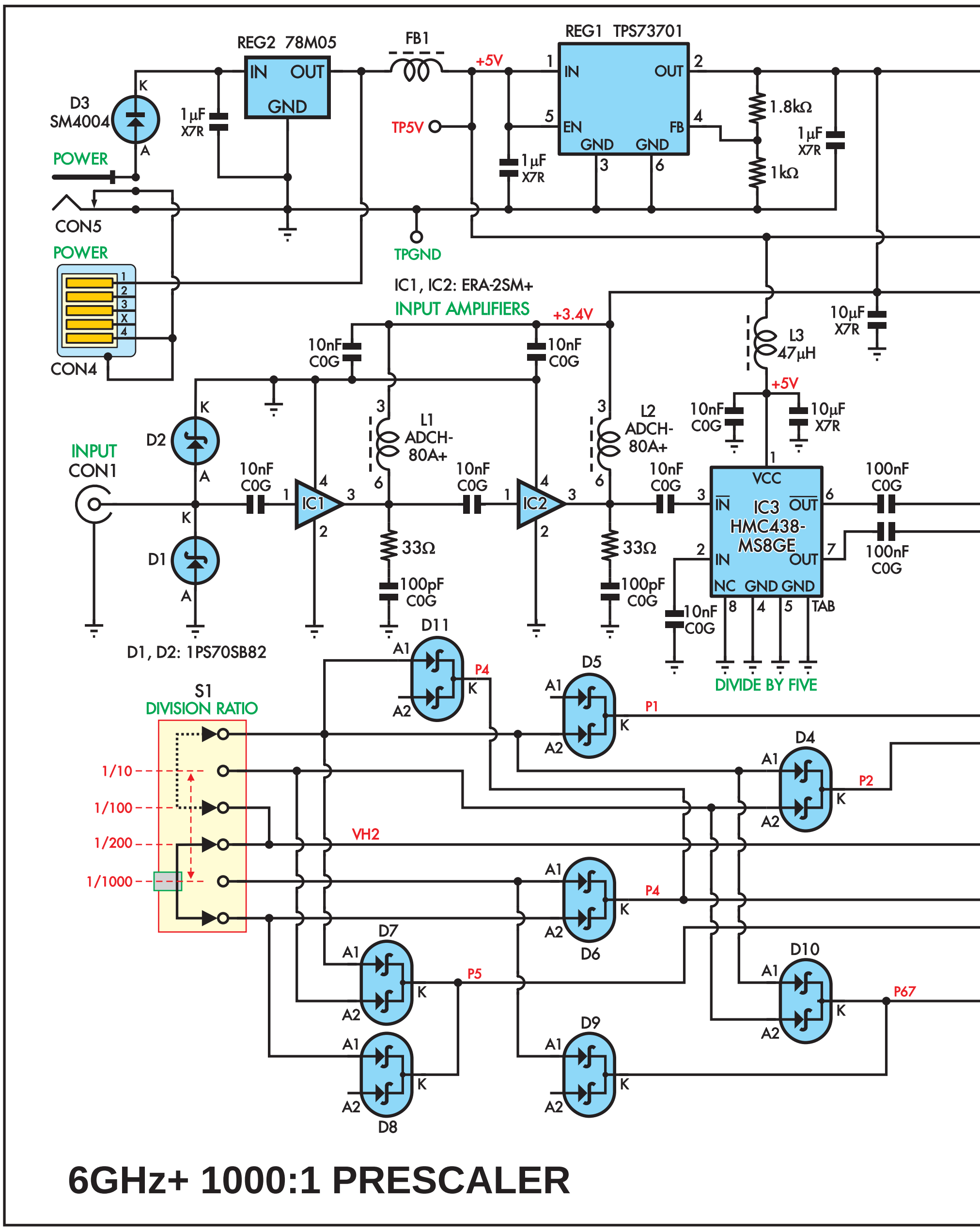
The lower signal limit shown in Fig.5, combined with the gain from IC1 and IC2, means that the theoretical sensitivity of the *RF Prescaler* is around –49dBm at 1GHz, which equates to an input signal of well under 1mV RMS. However, keep in mind that some of the input signal will be lost in the cabling and due to the 50Ω termination of the input, so in reality a 1mV signal would be marginal.

IC3 produces two output signals at one fifth its input frequency, with opposite phases, from pins 6 and 7. At low frequencies these are fairly square, although inevitably they become more sinewave-like at higher frequencies. These are coupled to another divider (IC4) via two 100nF capacitors. We're using higher value capacitors in these positions due to the lower frequency here compared to the input signal.

By extending the low frequency response of the unit, we reduce the need to constantly bypass the unit if you're measuring signals over a wide range of frequencies.

Programmable counter

IC4 is an eight-bit counter, counting from 0 to 255 (by default) and then going to zero again. If left in this default configuration (with most of the digital inputs open-circuit since they have internal pull-downs), the differential outputs C_{OUT} and $\overline{C_{OUT}}$ will produce pulses at a frequency 1/256th the input frequency ($256 = 2^8$). However, you can set IC4's division ratio to any value from 2 to 256. To do this, we set the states of input pins P_0-P_7 to an 8-bit digital value and pull the T_{CLD} input high. Every time the counter rolls over, rather than being reset to zero, it's loaded with the digital value from the P0-P7 pins.



Say we want an overall division ratio of 100. Since IC3 divides the input frequency by five, IC4 must divide the frequency by a factor of 20. To do this, we set P0-P7 to the binary value of 236 (256–20). Since counting now starts at 236, after 19 pulses, it reaches 255 (236+19) and so requires just one more pulse to roll over. Hence, it divides its input frequency by 20.

Selection of division ratios

As noted above, we're using a miniature 4-position horizontal slide switch (S1) to select the division ratios. This particular switch is a little unusual in that it has six pins and it works by bridging two of the pins, depending on the position of the switch, as depicted in the circuit diagram. For example, when in the 1/1000 position, the fourth and sixth pins are bridged.

We have arranged diodes D4-D11 so that in this position, the VH_2 voltage on the middle two pins of the switch (which we'll explain in more detail later) is applied to input pins P5 (via D8) and P4 (via D6), pulling those inputs high. Input P3 is permanently tied high. As a result, with P3, P4 and P5 high, the counter's initial binary value is 00111000 or 56 in decimal. Since 256 – 56 = 200 and 200 × 5 = 1000, we have the correct division ratio.

If you perform the same calculations for the other three switch positions, you will find that the pre-load counter values are 216 (256 – 40), 236 (256 – 20) and 254 (256 – 2).

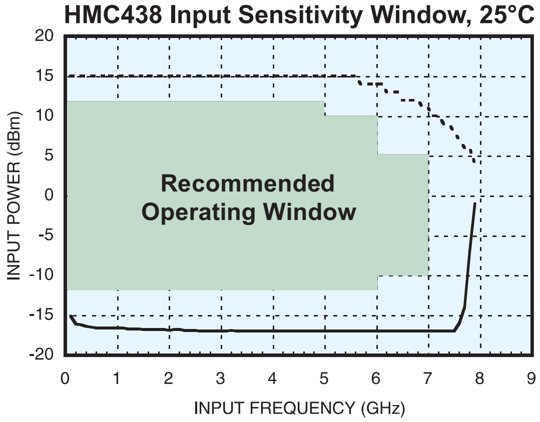


Fig.5: the recommended input power level for prescaler IC3 based on signal frequency. Keep in mind that IC3 is preceded by two amplifier stages for improved sensitivity.

ECL voltage levels

As mentioned earlier, IC4 is an ECL (emitter-coupled logic) device; a technology which has been used for decades for very high speed logic. ECL devices are bipolar transistors made from plain old doped silicon.

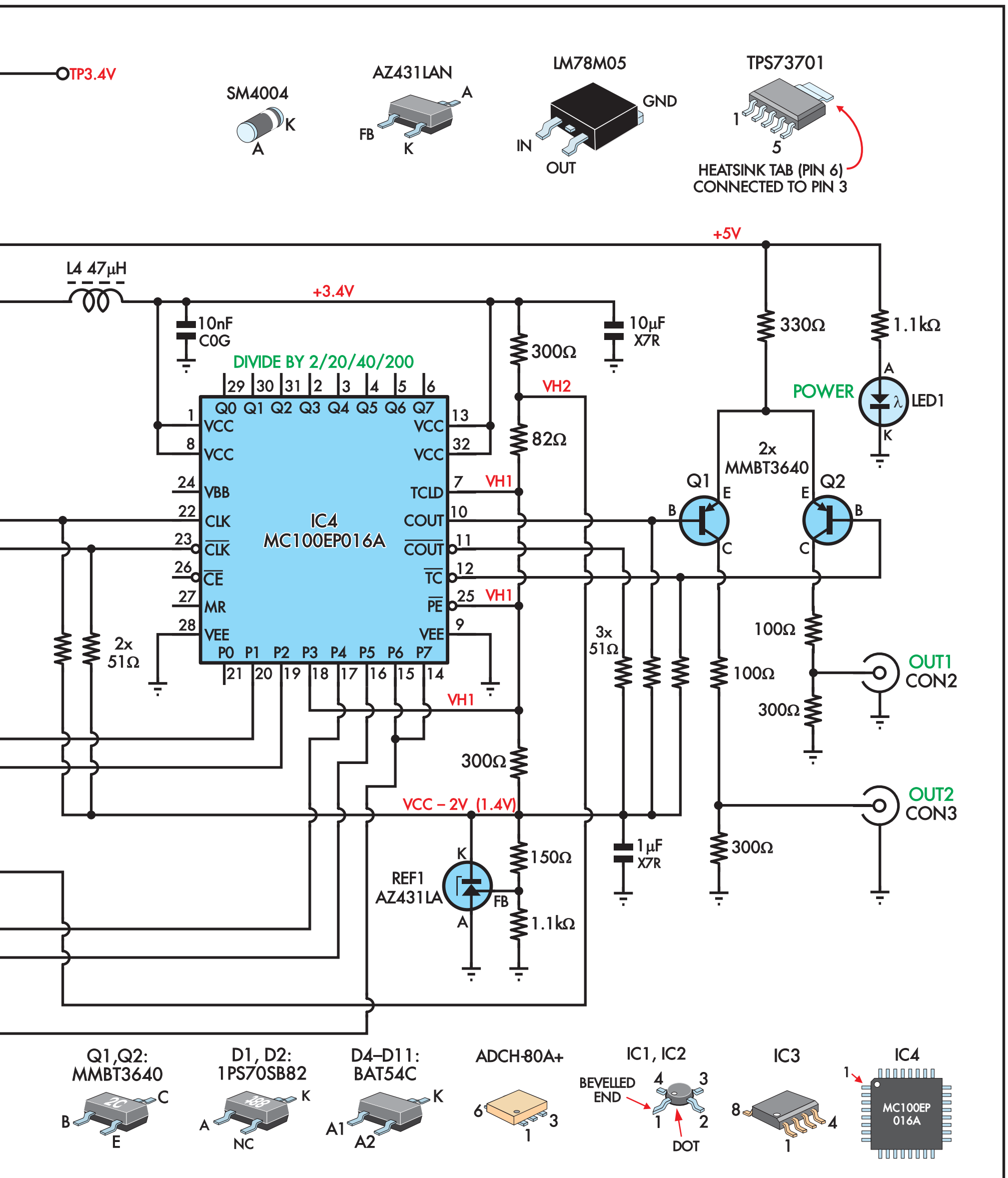


Fig.4: complete circuit for the *RF Prescaler*. The diode logic network comprising slide switch S1 and dual diodes D4-D11 configures IC4 for the selected division ratio.

Despite this, these transistors are arranged in such a way to allow operation at frequencies over 1GHz.

This is because the transistors are biased so that they are always conducting, with their conductance being varied to produce different digital states, rather than being switched on and off. In a sense, this means that they process digital information in an analogue manner. As a result, ECL input and output voltages swing over a much more limited range than CMOS or TTL.

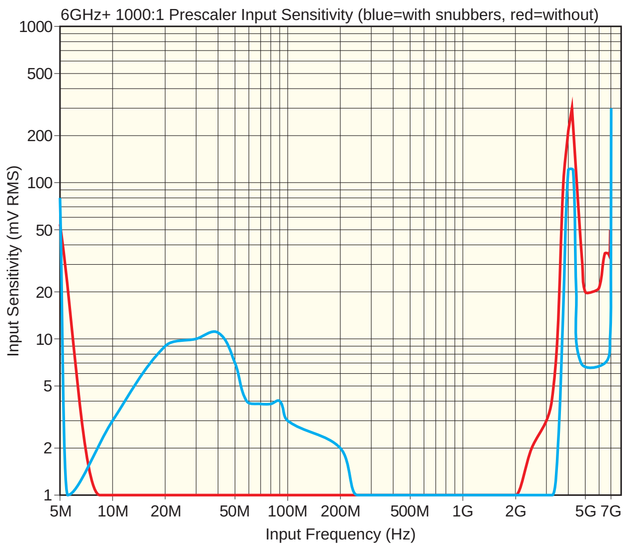
In the case of the MC100EP016A, the supply voltage is 3.0-3.6V and the average signal level is around 1V below this, ie, 2.0-2.6V, depending on the exact supply voltage. When a pin state changes between one and zero, typically its voltage will shift by around 0.7V. Assuming a 3.3V supply, a logic high level may be around 2.65V while a logic low would be around 1.95V.

Pin 24 on IC4 is labelled 'V_{BB}' and provides a reference voltage which is almost exactly halfway between the low and high stage voltages and may be used for comparison, to convert an ECL output to CMOS/TTL. We aren't using this pin though; we're using a different technique to produce the output signals, as will be explained later.

The somewhat unusual ECL levels do slightly complicate providing the correct input voltage levels for IC4. To achieve this, we have connected a resistive divider between the +3.4V rail and the 1.4V (VCC – 2V) rail to generate two additional voltage levels, VH_2 and VH_1. VH_2 is approximately +2.5V while VH_1 is approximately +2.3V.

VH_1 is therefore in the middle of the specified 'input high voltage' range for IC4 (with V_{CC}=3.4V) of 2.14-2.49V and so pins which are permanently tied high are held at this voltage, ie, TCLD (terminal count load; mentioned above), $\overline{PE}$ (the chip enable pin) and P3 (also mentioned above).

Fig.6: minimum input sensitivity for the *RF Prescaler*. Signal levels above this, up to about 1V RMS, should not be a problem. Below the level specified, it may operate with some jitter, or not at all. The blue curve is for the circuit as published, while the red curve shows its performance without the two snubber networks at the outputs of IC1 and IC2.



However, pins P1, P2 and P4-P7 are pulled high via a series of schottky diodes and switch S1, so VH_2 is connected to the anodes of these diodes rather than VH_1. This compensates for the voltage drop across the diodes, so that 2.3V is also applied to those pins when they are pulled high.

IC4's data sheet does not explain whether these inputs must be within the 'input high voltage' range, so we have played it safe and keep them within that range, rather than just tie them high (to +3.4V) and hope it works reliably.

The V_{CC}–2V (1.4V) rail which is used to derive VH_1 and VH_2 is generated by shunt regulator REF1. Its nominal voltage is 1.24V and the 150Ω/1.1kΩ resistive divider between its cathode, feedback input and anode sets the gain to 1.136 for an output of 1.41V (1.24V x 1.136).

This rail is also used to terminate the three main counter outputs of IC4 (COUT, $\overline{\text{COUT}}$ and $\overline{\text{TC}}$) via 51Ω resistors, in line with how the data sheet suggests they should be terminated to achieve the specified performance. REF1 can sink up to 100mA, which is more than enough for this application. The voltage across it is stabilised despite a high-frequency AC component to the current due to the 1µF bypass capacitor.

This same V_{CC}-2V rail is also used to DC-bias and terminate the CLK and $\overline{\text{CLK}}$ input signals for IC4 (at pins 22 and 23), via 51Ω resistors. Such low-value termination is done to ensure there's no overshoot or ringing overlaid on the signals from IC3, which might upset the operation of IC4.

Output stage

The differential output from IC4 is at pins 10 and 11 (C_{OUT} and $\overline{C_{OUT}}$) and being ECL outputs, these swing between about 1.95V and 2.65V. However, there is another output, $\overline{\text{TC}}$ at pin 12 which has a similar waveform to that at pin 11. We found its average DC voltage level more stable than that at pin 11, so we are using pins 10 and 12 as the differential outputs instead.

These are connected to a differential-to-single-ended conversion stage comprising 500MHz PNP transistors Q1 and Q2, which are arranged in a long-tailed pair. Since their emitters are joined together and supplied with current with a 330Ω fixed resistor from the 5V rail, the emitter voltage is determined by whichever base voltage is higher at the time. The bases of Q1 and Q2 are connected directly to the two outputs of IC4 mentioned above, pins 10 and 12.

Hence, whichever output is lower, the transistor it is driving is switched on harder, as it has a higher base-emitter voltage than the other. So when pin 10 of IC4 is lower, Q1 is switched on while Q2 is basically off, and when pin 12 is lower, Q2 is switched on while Q1 is basically off.

The collectors each have a total load resistance of 400Ω, arranged as a divider which reduces the collector signal voltage by 25% at output connectors CON2 and CON3, while providing an output impedance of 75Ω (ie, 100Ω in parallel with 300Ω). This results in an output voltage swing of around 2V peak-to-peak. However, when the output(s) are terminated with 50Ω or 75Ω, this is reduced to about 300mV peak-to-peak; sufficient to drive an external oscilloscope or frequency counter.

Power supply

For the power supply we recommend using a regulated 9V 500mA DC plug-pack, plugged into DC barrel connector CON5. This feeds 5V linear regulator REG1 via reverse-polarity protection diode D3, which in turn provides the 5V rail for IC3 and the output stage (Q1 and Q2) via a ferrite bead (FB1). FB1 prevents any high frequency modulation in the current draw of IC3 from radiating from the power supply lead.

The 5V rail is also applied to linear regulator REG2, which generates a 3.4V rail for IC1, IC2 and IC4. REG2 can either be an adjustable TPS73701 with 1.8kΩ and 1kΩ resistors connected to its feedback (FB) pin 4, as shown in Fig.4, or it can be a TPS73734 fixed 3.4V regulator.

If using the fixed regulator, omit the 1.8kΩ resistor and replace the 1kΩ resistor with a 10nF SMD capacitor, which gives it superior ripple rejection.

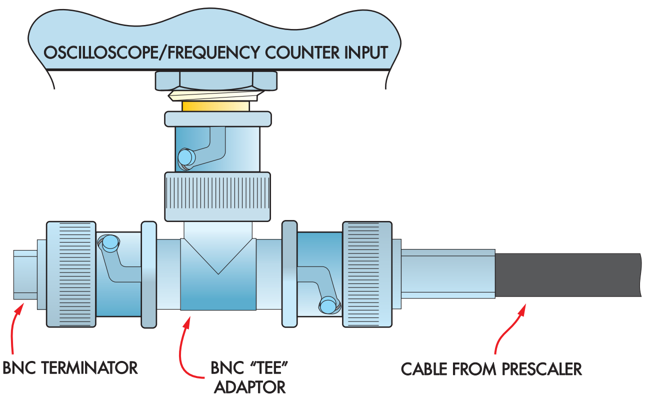
While we could have used a 3.3V fixed regulator, which is much more common than 3.4V, 3.4V is the ideal operating voltage for IC1 and IC2 (3.2-3.6V allowed) and is also suitable for IC4 (3.0-3.6V). Depending on tolerance, the output of a 3.3V regulator may be too low for proper operation of IC1 and IC2.

It's also possible to power the unit from a USB supply, via optional USB socket CON4. If both CON4 and CON5 are fitted, CON4 is automatically disconnected if a DC plug is inserted, by the switch integral to CON5.

While our unit successfully operated from a USB supply, because this supply is used to run IC3 directly, any significant high-frequency hash could interfere with its operation.

Since many USB chargers have quite poor regulation and high levels of hash, it's probably better to stick with the 9V supply option.

Fig.7: if you want to feed the output of the *RF Prescaler* to a device with a high input impedance (eg, 1MΩ or 10MΩ), here is the best way to do it. The signal must be terminated with a low impedance to get accurate results.



Apart from the four-position switch which selects the division ratio, there are no actual controls on the *RF Prescaler*. One edge has the SMA input socket (left), the division switch and the two BNC output sockets, one of which is 180° out of phase with the other. On the opposite side are the two power sockets – a 9V DC barrel socket (which we prefer) and a 5V micro USB socket (only one is used at any time) – if you only intend to use the 9V socket or the micro USB, the other can be left off the PCB, saving you a bit of drilling or filing.

Besides, drilling a round hole is a lot easier than cutting/filing a square hole!

Frequency limits

We've rated this *RF Prescaler* at '6GHz+' because as presented, it will definitely operate to at least 6GHz and probably up to 7GHz. The actual upper limit depends on the exact properties of ICs1-4 which are fitted to your board.

The signal first passes through amplifiers IC1 and IC2. These are rated to operate to 6GHz with a typical gain of 10.7dB at 6GHz; down from a peak of 16.4dB at lower frequencies (10-100MHz). Presumably, they will also provide gain for signal just above 6GHz but this is not specified in the data sheet. Our guess is they will operate to at least 6.5GHz with at least some gain and will probably pass signals to at least 7GHz.

IC3 can normally operate to at least 7.5GHz with no reduction in performance (see Fig.5) but sensitivity rapidly falls off above that and it's unlikely to work at 8GHz.

The data sheet for IC4 indicates that at standard room temperature, it will typically handle signals up to 1.4GHz and definitely up to 1.2GHz. That translates to 7GHz (1.4GHz × 5) typical input frequency and 6GHz (1.2GHz × 5) minimum guaranteed input frequency.

So you can see that with a bit of luck, the *RF Prescaler* should work up to 7GHz, albeit with reduced sensitivity.

Note that that you could replace the two ERA-2SM+ amplifiers with ERA-1SM+ amplifiers. These have a specified gain of 7.9dB at 6GHz and 8.2dB at 8GHz. However, do note that it's possible that IC4 won't handle these higher frequencies; after all, it's only guaranteed to work up to 1.2GHz. And the ERA-1SM+ has less gain at lower frequencies, for example, 12.1dB at 1GHz compared to 15.8dB for the ERA-2SM+. Hence our recommendation to use the ERA-2SM+ devices.

Construction

The *RF Prescaler* is built on a double-sided PCB which is available from the *EPE PCB SErvice*, coded 04112162, measuring 89 × 53.5mm. This is mounted in a diecast aluminium case. Almost all the components are SMDs, the exceptions being connectors CON-CON3 and CON5, switch S1 and power LED1. Use the PCB overlay diagram in Fig.8 as a guide during construction.

Start with IC4. You can use a standard soldering iron, as long as the tip is not too large, but we recommend that you purchase a small tube or syringe of flux paste and some solder wick if you don't already have some. Good light and a magnifier are also important.

Place a small amount of solder on one of the corner pads for IC4 and then orient the part on the board as shown in Fig.8. Pin 1 goes towards lower left – this should be indicated on the PCB silkscreen.

Once the IC is oriented correctly, heat the solder you applied to the corner pad and then carefully slide the IC into place and remove the heat. This process should take no more than a few seconds.

Now carefully check that the IC pins are centred on their pads. Check all four sides. Use magnification to make sure that all pins are properly centred on their pads. If not, re-heat the solder on that one pad and gently nudge the IC towards the correct position.

Repeat this process until you are happy that the IC is correctly located and check that its pin 1 is in the correct position before tack soldering the diagonally opposite pin.

Re-check that all the pins are correctly located; you can re-heat either solder joint at this point to make slight adjustments.

Now apply a thin layer of flux along all the IC pins and then apply solder to all the pins. Make sure you apply enough to get proper fillets. It's difficult to avoid bridging the pins at this point; what's most important is getting the solder to flow onto each pin and pad on the PCB.

Once all the pins have been soldered, apply another thin layer of flux paste and then use a piece of solder wick to remove any excess solder, especially where adjacent pins are bridged. Proceed carefully and re-apply flux paste if necessary.

When you have finished, clean off the flux residue (using either a proper flux solvent or ethyl alcohol/methylated spirits and a lint-free cloth) and examine the solder joints under good light and magnification to ensure they are all good and there are no more bridges left.

After soldering IC4, you can fit IC3 in the same manner. IC3 has smaller, more closely-spaced leads but there are only eight of them, on two sides of the IC. One additional thing you will have to take into consideration is that IC3 has a thermal pad on the underside and ideally, this should be soldered to the matching pad on the PCB.

If you have a hot air reflow system (lucky you!) this is quite easy, as it's just a matter of spreading some solder paste on the nine pads for this IC, putting it in position and then gently heating it until all the solder paste melts and reflows.

However, if you are just using a regular old soldering iron, you should spread a thin layer of solder paste on the large central pad, then drop the IC down into position and tack solder it in position.

After checking that its orientation and position are correct, solder the remaining leads using the same technique as for IC4. Then flip the board over and squirt some flux paste into the hole directly under IC3.

Melt some solder into this hole and heat it for several seconds. Remove heat and carefully check that IC3 is hot by quickly touching it with your finger.

This indicates that the solder has conducted enough heat through the hole to melt the solder paste you placed under it earlier.

If you're fitting microUSB connector CON4, do so now since its pins are hard to access once the other components are in place. This one is a little tricky because its pins are quite close together and despite the plastic locating posts, it's a little difficult to get the connector to sit in just the right position.

Start by putting a little flux paste on all the pads and pins for this device,

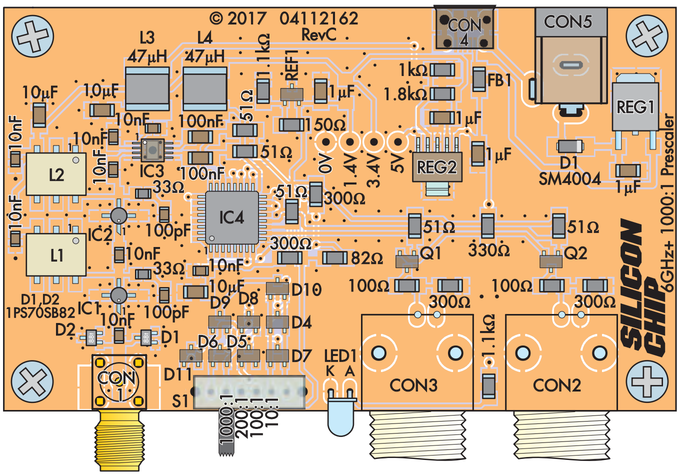


Fig.8: use this PCB overlay diagram as a guide to build the *RF Prescaler*. Start with IC4 and IC3 – these have the smallest pin spacings. Most of the remaining components are pretty easy to solder.

then drop it into place. Use a magnifying glass to check whether the pins are in the right location, then hold the device down with something heatproof (like a toothpick – not your finger!) and solder one of the large mounting lugs. This will take a few seconds as it will heat up the whole metal body while doing so.

Once you've formed a good solder joint on one of the mounting lugs, recheck that the signal pins are still located correctly. If they aren't, you will need to hold the socket with tweezers and nudge it into place while heating the solder.

You can then solder the remaining mounting lugs, followed by the signal pins and clean up any bridges between the pins using solder wick and a little extra flux paste. Use a magnifier to verify that all the signal pin solder joints are good before proceeding.

Remaining SMDs

The rest of the parts are quite easy to install as they have more widely spaced leads. Solder IC1 and IC2 next, making sure their 'pointy' pins are soldered to the pads marked for pin 1. Follow with L1 and L2, both of which are in six-pin packages. Their pin 1 dot should be oriented as shown in Fig.8.

Next on the list is REG1. This has one large pad and five small ones. The regulator itself has considerable thermal inertia, so spread a thin layer of flux paste on the large pad with a little extra paste on the smaller pads, drop REG1 in position and then tack solder one of the smaller pins (you can pre-tin the pad and heat it while sliding the part into place, if you like, as you did with IC4). You can clean these joints up with some additional flux paste and an application of solder wick.

Now for the large tab. Apply some solder to this tab and hold your iron in contact with both the regulator tab and PCB pad. You may need to hold it there for some time before the whole assembly heats up enough for the solder to flow down onto the board. Keep adding solder until the tab is covered and looks shiny, then remove the heat. Use a similar technique to fit REG2.

Inductors L3 and L4 are similarly quite large, so again, spread flux paste on each of their pads before soldering. You can then add some solder to one of the pads and slide the inductor into place while heating that solder.

Again, you may need to wait some time before the inductor heats up enough to slide fully into place and you can then add more solder until a nice, shiny fillet has formed. Let that cool down a little, then solder the opposite end, again waiting until it's hot enough to form a good joint (this should be quicker as both the inductor and PCB will retain significant heat).

The next components on the list are REF1, Q1, Q2 and diodes D4-D11. These are all in small 3-pin SOT-23 packages so don't get them mixed up. The eight diodes are all the same type. In each case, tack solder one pin, check that the pins are properly aligned, solder the other two pins and then refresh the initial pin. It's easier if you spread a little flux paste on the pads before soldering each part.

Now fit diodes D1 and D2, which are in similar but slightly smaller packages than D4-D11, followed by diode D3, which is in a two-pin rectangular or cylindrical package. Make sure its cathode stripe faces towards REG2 (indicated with a 'k' on the PCB). You can then fit all the ceramic capacitors and resistors to the board, as well as SMD ferrite bead FB1, where shown in Fig.8. Orientation is not critical for any of these.

Remember that if you're using a TPS73734 regulator, rather than the suggested TPS37301, you will need to omit the 1.8kΩ resistor and replace the 1kΩ resistor with a 10nF capacitor.

Through-hole components

With all the SMDs in place, you can now proceed to solder slide switch S1, SMA connector CON1, barrel connector CON5 (if being fitted) and BNC sockets CON2 and CON3. In each case, ensure the part is pushed down fully onto the PCB before soldering the pins. The larger metal connectors such as CON1 require quite a bit of heat to form good solder joints.

Note that the pads for CON1 are designed to allow either a right-angle or edge-mounting ('end launch') connector – we recommend using a right-angle connector like we did in our prototype, so that it lines up with BNC sockets CON2 and CON3.

Power indicator LED1 was not fitted to our prototype but we decided it would be handy and so have added it to the final version, located just to the left of output connectors CON2 and CON3. Bend its leads through 90° close to the base of the lens, so that the longest lead will go through the hole towards the right-hand side of the board, marked 'A' in Fig.8 and on the PCB. Solder it with around 6mm of lead length above the PCB, so that its lens lines up with CON1-CON3.

Initial testing and use

Ideally, you should connect an ammeter in series with the DC power supply the first time you fire up the *RF Prescaler*. Quiescent current should be close to 380mA (or 370mA on the 10:1 divider setting). Less than 350mA suggests that at least one device in the circuit is not getting sufficient voltage, while much more than 400mA possibly indicates a short circuit.

If the initial current drain is not in the range of 325-425mA, switch off immediately and carefully check the PCB for assembly faults, such as adjacent pins being bridged, bad solder joints or incorrectly placed or oriented components. Use good light, a magnifier and if necessary, clean flux (or other) residue off the board using methylated spirits or another similar solvent so that you can see it properly.

Assuming the current is in the right range, use a DMM to check the voltages at the three test points provided, labelled 1.4V, 3.4V and 5V. These are the voltages you should expect at each point. The 1.4V test point should be between 1.35V and 1.45V, the 3.4V test point between 3.35V and 3.45V, and the 5V test point around 4.75-5.25V (possibly slightly higher or lower if you're using the USB supply option).

If the 1.4V test point is off, that suggests a problem with REF1. If the 3.4V test point is off, you may have fitted incorrect divider resistors for REG2.

On our prototype, we used a TPS73701 (adjustable version of REG2) and found the 3.4V rail was a little low

Parts list – 1000:1 6GHz+ Prescaler

1 double-sided PCB, available from the *EPE PCB Service*, coded 04112162, 89 × 53.5mm
1 diecast aluminium case, 111 × 60 × 30mm
1 high-frequency SMD ferrite bead, 3216/1206 size (FB1)
2 Mini-Circuits ADCH-80A+ Wideband RF choke (L1,L2) (*available from* ***www.cseonline.com.au*** *or the* ***Silicon Chip Online Shop***)
2 47µH 6 × 6mm SMD inductors (L3,L4)
1 SMA right-angle through-hole or edge-mounting connector, 50Ω, >6GHz (CON1)
2 PCB-mount right-angle BNC sockets (CON2,CON3)
1 SMD microUSB socket (CON4) ***AND/OR***
1 PCB-mount 2.1mm or 2.5mm ID DC barrel socket (CON5)
1 C&K SK-14D01-G 6 PCB-mount right-angle SP4T micro slide switch (S1)
1 SMA male-to-BNC female adaptor (optional, for connecting BNC-equipped signal sources)
1 BNC T adaptor and 50Ω or 75Ω termination plug (optional, for driving high-impedance equipment)
1 9V DC regulated supply with plug to suit CON5 ***OR***
1 5V USB supply with Type-A to microUSB cable (see text)
4 M3 × 10mm pan-head machine screws and nuts
8 3mm ID 6mm OD 1mm thick Nylon washers
4 M3 Nylon nuts
4 small rubber feet (optional)

Semiconductors
2 Mini-Circuits ERA-2SM+ wideband RF amplifiers [Micro-X] (IC1,IC2) (*available from* ***www.cseonline.com.au*** *or the* ***Silicon Chip Online Shop***)
1 HMC438MS8GE 7GHz divide-by-five prescaler [MS8G] (IC3)
1 MC100EP016A 3.3V ECL 8-bit synchronous counter [LQFP-32] (IC4)
1 TPS73701DCQ (adjustable) or TPS73734DCQ (fixed) 1A low-dropout linear regulator (REG1)
1 78M05 5V 0.5A linear regulator [D-PAK] (REG2)
1 AZ431LANTR-G1DI 100mA 1.24V adjustable shunt reference [SOT-23] (REF1)
2 MMBT3640 12V 200mA 500MHz PNP transistors [SOT-23] (Q1,Q2)
1 3mm blue LED (LED1)
2 1PS70SB82 Schottky diodes [SOT-323/SC-70] (D1,D2)
1 S1G or equivalent 1A diode [SM-1/SMA] (D3)
8 BAT54C Schottky dual diodes [SOT-23] (D4-D11)

Capacitors (all SMD ceramic 3216/1206 size unless otherwise stated)
3 10µF 16V X7R
4 1µF 16V X7R
2 100nF 50V X7R
9 10nF 50V NP0/C0G, 2012/0805 size (one unused when REG1=TPS73701)
2 100pF 50V NP0/C0G, 2012/0805 size

Resistors (all SMD 3216/1206 size, 1%) * only required when REG1=TPS73734 ** may be required to trim REG1 output voltage

1 68kΩ**	1 30kΩ**	1 1.8kΩ*	2 1.1kΩ	1 1kΩ*	1 330Ω	4 300Ω
1 150Ω	2 100Ω	1 82Ω	5 51Ω	2 33Ω (2012/0805 size)		

at around 3.328V, presumably due to resistor tolerances. We solved this by soldering a 30kΩ resistor across the top of the 1kΩ resistor, bringing the 3.4V rail up to 3.399V.

We've added 30kΩ and 68kΩ resistors to the parts list. If your 3.4V rail is below 3.34V, solder the 30kΩ resistor in parallel with the 1kΩ resistor, while if it's between 3.34V and 3.37V, use the 68kΩ resistor instead. Between 3.37V and 3.5V should be OK. An output from REG1 above 3.5V is unlikely.

If you use the fixed version of REG2 (TPS73734) its output should be between 3.36 and 3.44V so it should not require any trimming.

Assuming the voltages seem OK, the next step is to hook the output(s) of the prescaler up to your frequency counter or scope. If this device has an option

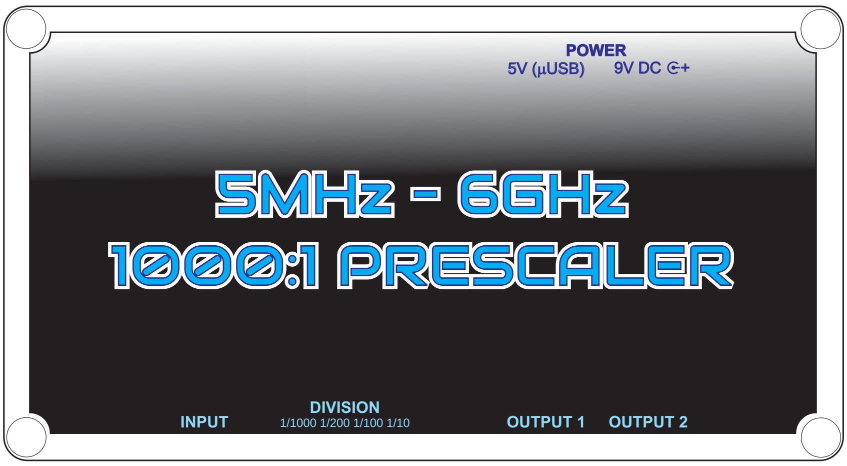


Fig.10: same-size artwork for the *RF Prescaler* front panel. There are no holes in the top panel to be drilled. We used only the inner portion of the artwork as you can see from our photos. You can photocopy this artwork without breaking copyright – or if you prefer, it can also be downloaded (as a PDF) from the *EPE* website.

for (or a fixed) 50Ω input impedance, select this. If your counter/scope only has a high impedance input, you will need to terminate the cable at its input using a 50Ω or 75Ω resistor.

Assuming this device has a BNC input, you can do this by connecting a BNC T adaptor to that input, with a termination plug on one end and the cable from the *RF Prescaler* on the other; see Fig.7.

You also need a signal source which can produce a signal of at least 5MHz (and ideally higher) into a 50Ω load. Connect this up to the *RF Prescaler's* input, power it up and check the reading from the output(s). Confirm that it is steady and in the expected range. Move switch S1 and check that the frequency reading is as expected on each setting; its left-most position is 1000:1 and right-most is 10:1.

Ensure that your signal generator can produce sufficient amplitude for correct operation, as shown in Fig.6, keeping in mind that the higher the frequency, the less signal you need for the *RF Prescaler* to operate. Note also that it will operate with signal levels a few dB below the sensitivity curve shown in Fig.6 with increasing jitter (and thus possibly decreasing accuracy in the reading) the further below the curve your signal is.

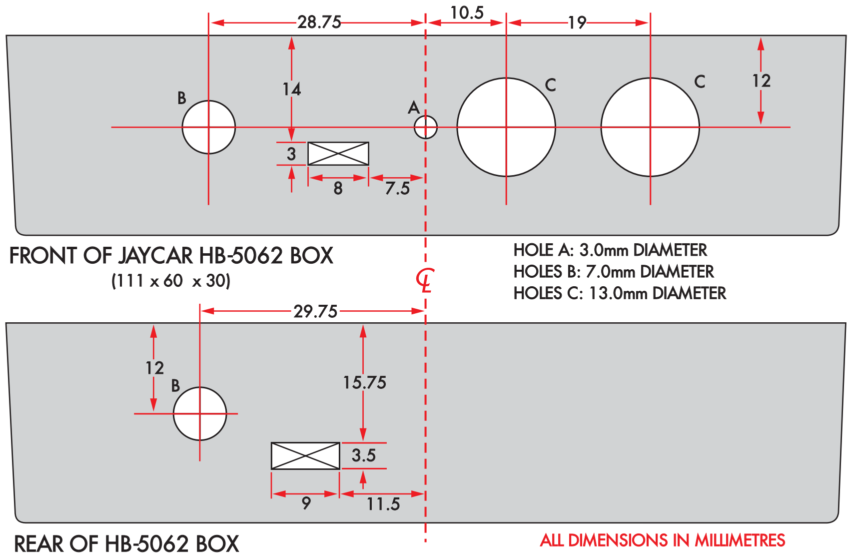


Fig.9: drilling detail for the diecast box. You don't need both the 7mm hole and the micro USB slot on the rear if you only intend to use one power source.

Putting it in a case

While we found the *RF Prescaler* operated reasonably well without a case, it's usually a good idea to shield RF equipment, both to prevent interference from affecting its operation and to prevent it from producing too much EMI which might affect other equipment.

Hence, our *RF Prescaler* is designed to fit in an inexpensive diecast aluminium case measuring 111 × 60 × 30mm (Jaycar HB5062). If you have a drill press and are reasonably experienced with machining aluminium, it should take you about one hour to install it in the case.

Start by printing out the drilling templates, shown in Fig.9 and also available for download as a PDF from the *EPE* website. Cut these out and glue/tape them onto the front and back of the case, centred as well as is possible.

Centre punch the holes and drill each one using a 3mm pilot hole. For the rectangular cut-out on the front panel, drill three 3mm holes inside the outline, one at either end and one in the centre.

The rectangular cut-out on the rear is only necessary if you're using a USB power supply. The rectangle shown is large enough to expose the microUSB connector; however, you will probably have to expand it considerably to get the plug to fit in. Alternative, if using a DC plugpack (as recommended), you can drill the adjacent hole instead.

Once each pilot hole has been drilled, use either a stepped drill, series of larger drill bits or tapered reamer to enlarge each hole to its final size. File any rectangular cut-outs flat and then enlarge them to size.

Make sure each hole is clean (ie, no swarf) and get rid of all the aluminium shavings, then remove the nuts and washers from the BNC connectors and test fit the PCB in the case. You will need to angle it in. The front panel holes are slightly oversize to give you enough room to do so.

Don't force it in if it won't go in easily; if you do, you may not be able to get it out! Simply enlarge the holes slightly and it should pop in with only modest force and you can then drop it down to be parallel with the base. We suggest that you put switch S1 in one of the centre positions initially, then once the PCB is in the case, make sure the slot is wide enough to allow all four positions to be used.

Make sure that you check that the rear panel hole(s) are large enough to make a good power supply connection to the PCB. Most barrel plugs should be long enough to fit through the hole and into the connector. If yours isn't, you may need to cut it off and solder a longer one onto the plugpack.

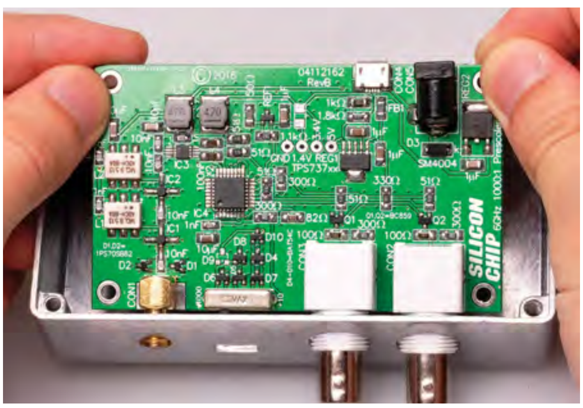
With the PCB in the case, you can now use it as a drilling template to drill four 3mm holes in the base. Remove the PCB by lifting the rear and then pulling it out, then clean out the aluminium dust.

Now, feed a 10mm machine screw up through one of the holes in the base and place two of the 1mm-thick nylon washers over its shaft, then screw on a nylon nut until the screw thread is just about poking through the nut. Repeat for the other three holes. If you're using screw-on rubber feet, you should pass the 10mm machine screws up through the feet before feeding them into the case.

If you lift the case up, the screws should drop down, leaving just the two nylon washers and nut sitting on the bottom of the case in each corner.

This should give you enough room to lever the PCB back in. Press down on one corner of the PCB and rotate that screw clockwise until its shaft is just poking through the PCB, then hold an M3 nut down on the shaft and continue tightening until the screw has gone all the way into the base and the nut is holding the PCB down.

Repeat for all four corners. You can now place the washers back over the BNC connectors and screw the nuts back on.



Fitting the completed PCB into the case is very much a 'shoehorn' affair, but it does fit! Don't force it – a bit of judicious 'jiggling' should get it in place.

Win a Microchip Curiosity HPC Development Board

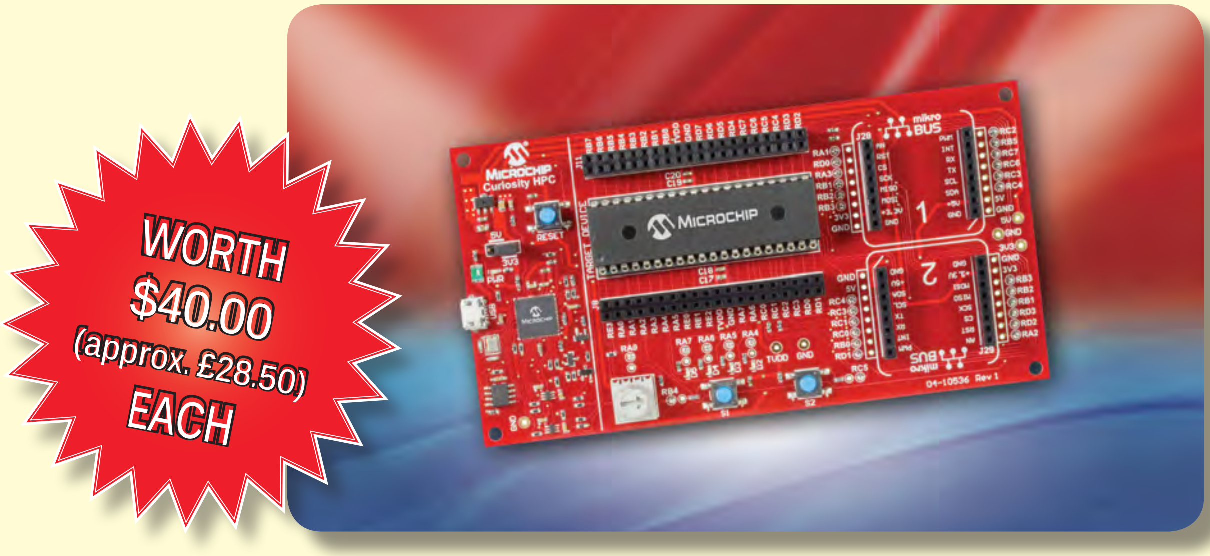
EVERYDAY PRACTICAL ELECTRONICS is offering its readers the chance to win a Win a Microchip Curiosity HPC Development Board (DM164136).

Curiosity HPC is the perfect platform to harness the power of modern 8-bit PIC microcontrollers. Its layout and external connections offer unparalleled access to the Core Independent Peripherals (CIPs) available on many newer 8-bit PIC MCUs. These CIPs enable the user to integrate various system functions onto a single MCU, simplifying the design and keeping system power consumption and BOM costs low.

This board provides flexibility for experimentation through an application header with ground (GND) and supply voltage (VDD) connections. It also includes a set of indicator LEDs, push button switches, and a variable potentiometer. Additionally, it features two mikroBUS headers to accommodate a variety of plug-in Click boards that can be used in application development. All connections to the mikroBUS headers, LEDs, switches and potentiometer are labelled with the microcontroller port name for ease of programming. The full pin breakout of the microcontroller is provided to expand the flexibility of the Curiosity HPC Development Board.

Curiosity HPC is fully compatible with MPLAB Code Configurator and MPLAB X v3.05 or later.

The Curiosity HPC Development Board accommodates 40- and 28-pin 8-bit microcontrollers.



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Micromite BackPack V2

with Touchscreen LCD and onboard programmer

By Geoff Graham

The Micromite LCD BackPack described in the May 2017 issue has been a very popular project. This revised version incorporates the Microbridge described in this issue. It adds a USB interface and the ability to program/reprogram the PIC32 chip while it's onboard. And the BackPack V2 also adds software control over the LCD backlight.

The Micromite *LCD BackPack* has been a huge hit since it was introduced in May last year. For those who missed it, the *BackPack* combines the Micromite, which is a low-cost, high-performance microcontroller programmed in BASIC, with an equally low-cost LCD touchscreen.

Together, the pair make a potent combination, allowing you to easily design a gadget with an advanced user interface. We have published quite a few examples of this, for example, the *DDS Signal Generator* in last month's issue of *EPE* – see April 2018.

While the original *Micromite LCD BackPack* was easy to build, it did require you to use an external USB/serial converter so that you could load and run programs.

You also needed a PIC32 programmer to load and update the MMBasic firmware in the Micromite, and many people felt that the cost of a genuine PICkit 3 programmer from Microchip was too expensive.

This new design includes both the USB/serial interface and PIC32 programming capability in a single additional chip, dubbed the *Microbridge* – see the separate article describing its operation in this issue.

Because the *Microbridge* is so cheap, it has been designed to be a permanent part of the *Micromite BackPack V2*. So now you can update the firmware in the Micromite and edit your BASIC program without any extra hardware.

We have also included the ability to control the LCD backlight brightness from within the BASIC program running on the Micromite.

This requires just four additional components plus the use of an extra I/O pin on the Micromite. These components are optional; you can either include them or use the original brightness control arrangement with a trimpot (keeping the PWM pin free for other uses).

Apart from the above additions, this new version of the *Micromite LCD BackPack* is exactly the same as the original. It is programmed in the same way, the I/O pins are the same and it will happily run programs written for the original version. It's the same basic formula but easier to use.

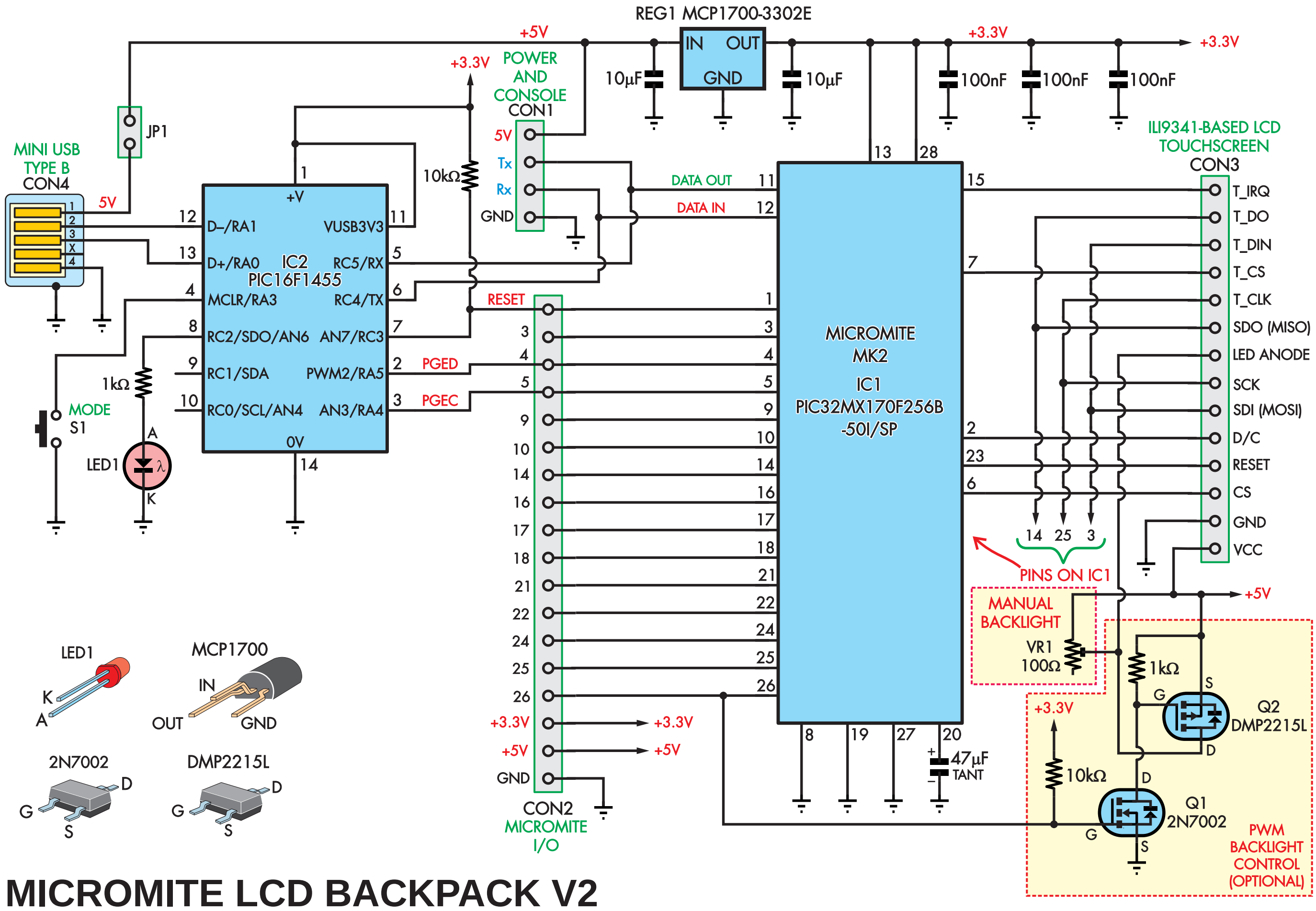
Circuit details

Fig.1 shows the complete circuit for the revised *Micromite LCD BackPack*, incorporating the *Microbridge*. IC2 is a Microchip PIC16F1455 microcontroller, which is both a USB/serial converter and a PIC32 programmer – the standalone *Microbridge* article (see page 28) describes its function in more detail.

When running as a USB/serial converter, pin 5 on the PIC16F1455 receives data (ie, data from the Micromite to the PC USB interface) and pin 6 transmits data (from the PC USB interface to the Micromite). These signals also run to the edge pins for the console connection (CON1) in case you build this PCB but for some reason do not plug the *Microbridge* IC (IC2) into its socket. In this case, you can use an external USB/serial converter.

The PIC32 programming interface from the Microbridge is on pins 7, 2 and 3 of IC2. These provide the reset function, program data and clock signals respectively. These connect to pins 1, 4 and 5 on the Micromite (IC1). The programming output on the *Microbridge* is only active when it is in programming mode, so the *Microbridge* does not interfere with the Micromite when it is using pins 4 and 5 as general purpose I/O pins.

As described in the *Microbridge* article, switch S1 is used to select programming mode and LED1 indicates the mode (lit solid when in programming mode).



MICROMITE LCD BACKPACK V2

Fig.1: complete circuit of the *BackPack V2*, incorporating the *Microbridge* (IC2) which acts as both a USB/serial converter and PIC32 programmer. Micromite chip IC1 runs the show, while REG1 supplies both ICs with a regulated 3.3V. IC1 has an internal 'core' regulator to provide itself with 1.8V which is filtered by the external 47µF tantalum or ceramic capacitor.

CON2 is the main I/O connector for the Micromite and is designed so that it can plug into a solderless breadboard for prototyping. The connector also makes it easy to add a third PCB to the *LCD BackPack* 'stack', which can carry circuitry specific to your application (such as amplifiers or relay drivers). This connector is wired identically to the original *BackPack*.

The Micromite communicates with the LCD panel using an SPI interface where pins 3 and 14 (on the Micromite) carry data to/from the LCD, while pin 25 provides the clock signal. When the Micromite pulls pin 6 low, it is communicating with the LCD panel, and when pin 7 is pulled low, the Micromite will be communicating with the touch controller on the display panel.

The 28-pin Micromite has only one SPI port and so pins 3, 14 and 25 (SPI data and clock) are also made available on CON2 so that you can also use this SPI serial channel to communicate with external devices.

Backlight control

For controlling the brightness of the LCD's backlight you have two choices. The first is to fit MOSFETs Q1 and Q2 to the PCB, along with their associated resistors (this area is marked with a box on the PCB). When you do this, PWM output 2A on the Micromite is used to control the backlight brightness from within your program. This is described in more detail later.

Alternatively, as with the original *BackPack* you can fit VR1, which is a 100Ω trimpot. This is in series with the power to the backlight LEDs so it limits the current drawn by them and therefore sets the brightness. Note that you should install one set of components or the other (not both).

In both cases, the LCD panel has a 3.9Ω resistor in series with the backlight so you will not burn out the backlight if you set the PWM output to 100% or wind VR1 all the way around to zero ohms.

The power supply is derived from either the 5V connector pin on CON1, or if JP1 is installed, from USB connector CON4. Powering the *Micromite LCD BackPack* from USB power is handy during program development, but for an embedded controller application, you would normally remove the jumper from JP1 and supply 5V power via CON1. Note that you should not try to power the *BackPack* from both CON1 and USB as you could cause damage to the USB interface on your computer.

The 3.3V power supply for both the Micromite and the *Microbridge* is provided by REG1, which is a fixed output regulator with a low dropout voltage suitable for use with USB power supplies. This supply is also made available on CON2 so you can use it for powering external circuits (to a maximum of 150mA).

Sourcing the LCD panel

The ILI9341-based LCD panel used in the *Micromite LCD BackPack* comes in three sizes: 2.2-inch, 2.4-inch or 2.8-inch diagonal.

The PCB for the *Micromite LCD BackPack V2* is designed to suit the mounting holes for the 2.8-inch version; however, compatible displays of any of these three sizes will plug into the PCB and will work perfectly. So your only issue with using a 2.2-inch or 2.4-inch display will be that you will need to use some other physical mounting arrangement.

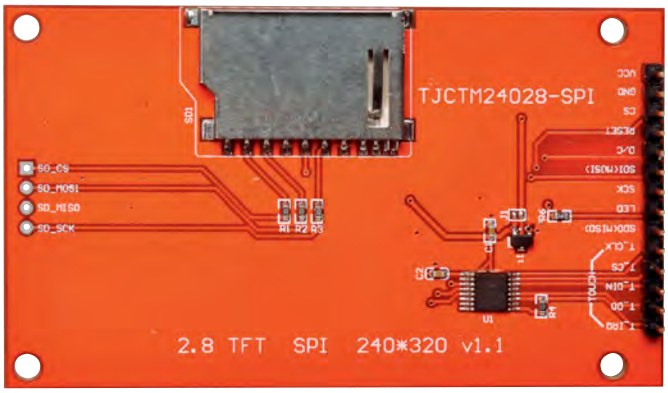
These displays also have an SD card socket, but it's not Micromite supported due to memory limitations.

The best place to find a suitable display is AliExpress or eBay, but

Firmware updates

For firmware updates and manual please check the author's website at: **geoffg.net/micromite.html**

You should also check out the Back Shed forum at: **www.theback-shed.com/forum/Microcontrollers** where there are many Micromite enthusiasts who are happy to help beginners.



The underside of the 2.8-inch ILI9341-based LCD panel we used in the *Micromite BackPack V2*. On the other side of the PCB to the top right of the LCD screen are the letters 2812C-SZ, which may prove useful when searching for this module.

other online markets also have them as well as some online retailers. There are many variations on offer, so make sure the display you purchase matches the photographs in this article. This is important; the Micromite has been extensively tested with the photographed display so you can be sure that it will work. Also ensure the touch controller is installed.

Other features to look out for in a compatible display are an orange PCB, a resolution of 320 × 240 pixels and an SPI interface. Often, the description will emphasise that the display is for use with the Arduino, but that is not relevant; it will work just as well with the Micromite. On eBay, the best way to find a suitable display is to search for the phrase 'ILI9341 LCD'. You should find many displays from US$7.00 upwards.

If you don't want to deal with any of that, then we recommend you purchase a kit from **micromite.org** – our preferred supplier for all things Micromite. The kit includes the LCD touchscreen, PCB, programmed microcontrollers and all the other bits you need to build the *BackPack V2* (apart from the acrylic lid)

Construction

Refer to the PCB overlay diagram, Fig.2. As usual, start construction with the low profile components such as resistors and work your way up to the bigger items such as the connectors.

Begin with the USB socket as this is the only required SMD component. Match the two small plastic pegs on the connector with the corresponding holes on the PCB and then solder the connector's mounting lugs using plenty of solder for strength.

Finally, using a fine point soldering iron tip, solder the signal pins. Examine the pin solder joints carefully under good light with magnification and clean up any bridges with solder wick and a little flux paste.

If you are installing the backlight PWM control components, you should mount Q1 and Q2 next as they are also surface mount types. They are not hard to solder as their pin spacing is quite wide. Don't get them mixed up as they look almost identical.

We recommend using a socket for both IC1 and IC2 as that will enable you to swap out the chips if you suspect that you have damaged one or both. The 14-pin female connector used for CON3 (the LCD panel) is difficult to source so unless you've purchased a kit, the best approach is to cut down a longer header to size and then use a file to smooth the rough edge so that it looks presentable.

The 10µF and 47µF tantalum capacitors are polarised (the longer lead is positive) so make sure that they are oriented according to the silk screen on the PCB. The 47µF capacitor is particularly critical and must be a tantalum or ceramic type, not electrolytic. Rather than using tantalum capacitors, we prefer to use SMD ceramic types with an X5R dielectric. In this case, you can use 10µF 6.3V capacitors in all three locations. They tend to be more reliable than tantalums, but are not as easy to obtain.

When soldering the pin headers for CON1 (power) and CON2 (input/output), remember that the headers should be mounted on the underside of the board, as illustrated in the photos. Don't mistakenly mount them on the top of the board because they will then be impossible to reach when an LCD panel is attached.

Before you plug the microcontrollers into their sockets, it is prudent to apply power and check that 3.3V is on the correct pins of IC1 and IC2, and 5V is on the correct pin of CON3. With that check made, remove power and plug in both microcontrollers and the LCD panel.

If you have a blank PIC16F1455 microcontroller, it should be programmed with the latest *Microbridge* firmware (**2410417A.HEX**), which can be downloaded from the *EPE* website. This can also be done using another Micromite and a 9V battery; see the *Microbridge* article for details on how to do this.

The *BackPack* PCB and the LCD panel can then be fastened together on all four corners with 12mm tapped spacers and M3 machine screws. Be careful when handling the LCD panel. The ILI9341 controller is sensitive to static electricity and can be easily destroyed with careless handling. Make sure that you are grounded when handling the display and avoid touching the connecting pins.

Programming the PIC32

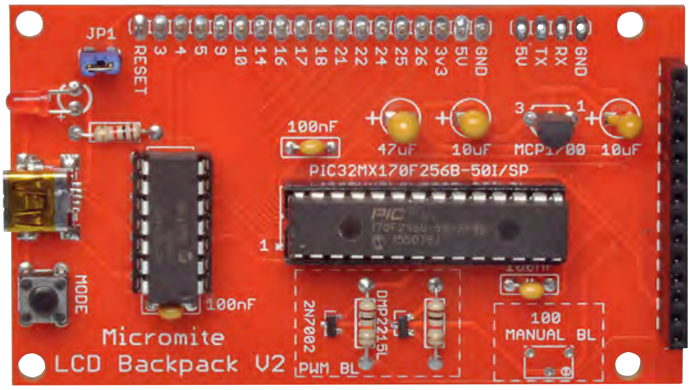
If you have a blank PIC32 chip, this needs to be programmed with the Micromite firmware via the *Microbridge*. This procedure is covered in detail in the *Microbridge* article so we will only provide an abbreviated description here.

The first step is to get the *Microbridge* working as a USB/serial bridge. This involves installing the correct drivers (available from **www.microchip.com/wwwproducts/en/MCP2200**) and launching a terminal emulator and connecting to the COM port created by the *Microbridge*. You can verify that everything is working correctly by typing characters into the terminal emulator and checking that LED1 on the *BackPack* flashes with each keystroke.

Now close the terminal emulator. This is important, as the programming operation will fail if it is still open. You need a Windows computer for the next step. Run the program **pic32prog** (available for download from the author's website) in a command prompt box with the command line:

```
pic32prog -d ascii:comxx yyyy.hex
```

Where xx is the COM port number created by Windows for the *Microbridge* and **yyyy.hex** is the file containing the latest Micromite firmware. For example, if your *Microbridge* was allocated the virtual serial port of COM6 and the file that you wanted to program was **Micromite_5.04.08.hex**, the command line that you should use would be:



The *Micromite LCD BackPack V2* includes the *Microbridge* (the 14-pin chip at left) which incorporates a USB/serial converter and a PIC32 programmer. You can also control the LCD backlight brightness via the BASIC program running on the Micromite. This uses four components that can be seen below IC1. Note, this is an early prototype and the final PCB differs slightly (it includes an extra 10kΩ resistor above IC2).

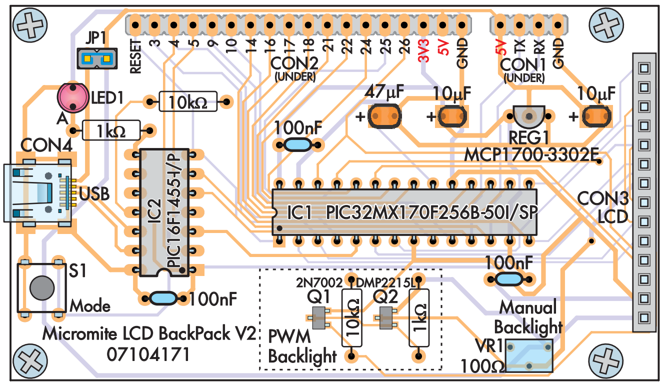


Fig.2: follow this overlay diagram to build the *Micromite LCD BackPack V2*. CON4 is the only required SMD component; SMD ceramic capacitors can optionally be used in place of the tantalum types for better reliability. If fitting Q1 and Q2, be sure to also install the two associated resistors and leave VR1 out. Note that CON1 and CON2 are fitted to the underside of the board.

pic32prog -d ascii:com6 Micromite_5.04.08.hex

When you press Enter, **pic32prog** will automatically run through the programming sequence and then return to USB/serial mode. You can then launch your terminal emulator and when you press return you should see the Micromite command prompt (a greater than symbol '>').

Fault finding

Your *BackPack* should work first time, but if it does not, the first thing to do is check that the correct power voltages are on the IC1 and IC2 sockets and CON3 (the LCD connector). Then check the 5V current drain for the full module, including the LCD; it should range from 100mA to 200mA, depending on the setting of the backlight. If it is substantially lower than this, check that the PIC32 and the LCD are correctly seated in their sockets.

With the LCD removed, the current drain should be about 30mA. If it is a lot less than this, it indicates that the PIC32 processor has not started up and in that case, the 47µF capacitor is the most likely culprit. It must be a tantalum or multilayer ceramic type; not an electrolytic.

If the current drain is correct, check that the *Microbridge* is working correctly. Does your PC recognise it as a valid USB device? Do you have the correct driver installed? Do you have your terminal emulator configured correctly?

You can check the *Microbridge's* operation by typing characters into your terminal emulator and watching for the LED to flash as they are received by the *Microbridge*.

Configuring the Micromite

The next step is to configure the Micromite for the LCD panel. To do this, type the following line at the command prompt (via the USB/serial connection and your terminal emulator software) and hit the enter key:

OPTION LCDPANEL ILI9341, L, 2, 23, 6

This tells the Micromite that the LCD panel is connected and which I/O pins are used for critical signals such as reset and device select. This option only needs to be entered once because the Micromite will store the setting in internal non-volatile memory and will automatically recall it whenever power is applied.

Following this command, the Micromite will initialise the display (which should go dark) and return to the command prompt. You can test the display by entering the following at the command prompt:

GUI TEST LCDPANEL

This will cause the Micromite to draw a series of rapidly overlapping coloured circles on the display as shown in the photo overleaf. This animated test will continue until you press a key on the console's keyboard and MMBasic will then return to the command prompt. To configure the touch feature, enter the following at the command prompt:

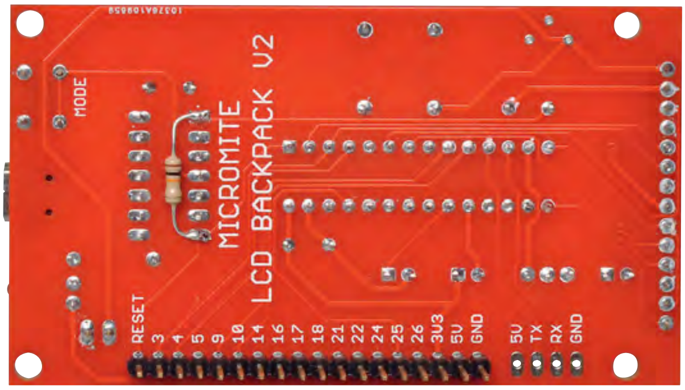
OPTION TOUCH 7, 15

This allocates the I/O pins for the touch controller and initialises it. This option is also stored in non-volatile memory and is automatically applied on power-up. Before you can use the touch facility, you need to calibrate it. This is done with the following command:

GUI CALIBRATE

This will cause MMBasic to draw a target in the upper left-hand corner of the screen. Using a pointy but blunt (ie, not too pointy) object, such as a toothpick, press on the exact centre of the target. After a second, the target will disappear and when you lift your implement another target will appear at upper right.

Continue pressing on the targets in this fashion until you have calibrated all four corners of the screen. The message



The underside of the prototype *LCD BackPack V2* contains the pin connections for the Micromite. Note that the 10kΩ resistor soldered between pins 1 and 7 of the PIC16F1455 is soldered through-hole on the top layer of the final PCB.

This is what the screen looks like when running 'GUI TEST LCDPANEL' as it draws a series of coloured circles on top of one another.

'Done. No errors' should be displayed on the console. You also might get a message indicating that the calibration was inaccurate and in that case you should repeat it, taking more care to press steadily on the centre of each target.

As before, these calibration details are saved in non-volatile memory and will be re-applied at power up. You can now test the touch facility with the command:

```
GUI TEST TOUCH
```

This will clear the screen and when you touch it, pixels will be illuminated at the touch point. This enables you to test the accuracy of the calibration. Pressing any key in the console will terminate the test.

Using the Microbridge

Using the *Microbridge* interface is quite easy. If you have identified the COM number allocated by your operating system, you can enter this into the set-up of your terminal emulator (we recommend Tera Term for Windows). The *Microbridge* defaults to a speed of 38,400 baud, so your terminal emulator will need to be set to a value of 38,400 baud to match the default speed used by the Micromite's console.

You can change the interface to a higher speed, which makes program loading faster and more convenient. For example, at 230,400 baud the built in Micromite editor (the EDIT command) is blazingly fast. To make the change, you need to set the interface speed on the Micromite and then in your terminal emulator. First, change the speed of the Micromite by issuing the following command at the command prompt:

```
OPTION BAUDRATE 230400
```

The Micromite will immediately switch to this speed so you will see some junk characters in your terminal emulator window. You then need to re-configure your terminal emulator for 230,400 baud. Press Enter and you should see the MMBasic command prompt ('>'). Both the terminal emulator and the Micromite will remember this new speed so you do not need to set it again.

If you configure the Micromite to some other baud rate and forget what it is, you may be stuck with a Micromite that you cannot communicate with. If that happens, you can restore the Micromite to its original defaults using the *Microbridge*.

The reset can be performed by pressing the mode switch on the *Microbridge* for two or more seconds, while simultaneously sending a continuous stream of exclamation marks at 38,400 baud, via your terminal emulator. Then release the mode switch while still sending exclamation marks for another two or more seconds. This causes the LED to flash and the MCLR line is briefly driven low to cause the reset.

Parts list

1 double-sided PCB, available from the *EPE PCB Service*, coded 07104171, 86mm × 50mm
1 ILI9341-based touchscreen LCD panel, 320 × 240 pixels, 2.8-inch diagonal (2.2 or 2.4-inch displays need special mounting)
1 PCB-mount SPST momentary tactile pushbutton (S1)
1 100Ω 0.5W vertical side-adjust trimpot (only fit if Q1 and Q2 are omitted)
1 28-pin narrow low-profile DIL IC socket (for IC1)
1 14-pin low-profile DIL IC socket (for IC2)
1 2-pin male header, 2.54mm pitch and jumper shunt (JP1)
1 4-pin male header, 2.54mm pitch (CON1)
1 18-pin male header, 2.54mm pitch (CON2)
1 14-pin female header socket, 2.54mm pitch (CON3)
1 mini Type-B USB 2.0 socket, SMD mounting (CON4)
4 M3 × 12mm tapped spacers
4 M3 × 6mm pan-head machine screws
4 M3 × 8mm pan-head machine screws
4 nylon washers, 3mm ID, 6mm OD, 1mm thick
1 laser-cut lid (optional)

Semiconductors
1 PIC32MX170F256B-50I/SP microcontroller – a PIC32MX170F256B-I/SP can be used but will be limited to 40MHz
1 PIC16F1455-I/P microcontroller programmed with Microbridge firmware (IC2) – the PIC16LF1455-I/P and PIC16(L)F1454-I/P are also suitable
1 MCP1700-3302E/TO 3.3V linear regulator (REG1)
1 3mm red LED (LED1)
1 2N7002 N-channel MOSFET, SOT-23 package (Q1) (optional, for PWM-controlled LCD backlight)
1 DMP2215L P-channel MOSFET, SOT-23 package (Q2) (optional, for PWM-controlled LCD backlight)

Capacitors
3 100nF multi-layer ceramic
2 10µF 16V tantalum or SMD ceramic, X5R, 3216 (1206) size
1 47µF 16V tantalum or 10µF SMD ceramic, X5R, 3216 (1206) size

Resistors (all 0.25W, 5%)
2 10kΩ (1 optional, for PWM-controlled LCD backlight)
2 1kΩ (1 optional, for PWM-controlled LCD backlight)

This will completely restore the Micromite to its initial configuration of 38,400 baud and erase any program and options held in memory. As a result, you will need to re-configure the Micromite for the LCD panel as described earlier.

Backlight control

If you installed the 100Ω trimpot for manual backlight control, the brightness adjustment is as simple as tweaking VR1 to your preference.

If you installed the components for the PWM-controlled backlight (ie, Q1, Q2 and the two associated resistors), the brightness is controlled via the PWM command in MMBasic. By default, the backlight will be at full brightness but it can be controlled with the following command:

```
PWM 2, 250, xx
```

where 'xx' is the percentage of full brightness required. This can range from 0 to 100. For example, a brightness of 75% is a good compromise between visibility and power consumption and this can be set with the following command:

```
PWM 2, 250, 75
```

Within a program, you can get a nice fade from full brightness to black by using the following program fragment:

```
FOR i = 100 to 0 STEP -1
  PWM 2, 250, i
  PAUSE 4
NEXT i
```

The PWM output used for the backlight control appears on pin 26, so this pin is not available for general I/O if you installed the components for the programmed controlled backlight.

Interfacing with other circuitry

The *Micromite LCD Backpack* interfaces to the world using CON2, the main I/O connector. This is designed so that you can plug it into a solderless breadboard or connect to a third board mounted on the back of the *BackPack* (eg, see the *Touchscreen Voltage/Current Reference* project in the October and December 2017 issues). The silk screen on the PCB identifies each pin on the connector. The GND, 5V and 3.3V pins can be used to power your external interface circuitry.

The maximum current that can be drawn from the 3.3V pin is 150mA, while the maximum 5V load will depend on your 5V supply. The RESET pin is normally at 3.3V, pulled up by the onboard 10kΩ resistor, and if you pull it low the Micromite will reset.

The other I/O pins connect directly to the Micromite and are marked with the Micromite pin number. You should refer to the *Micromite User Manual* (available for download from the author's website **http://geoffg.net/micromite.html**) for details of what you can do with each pin.

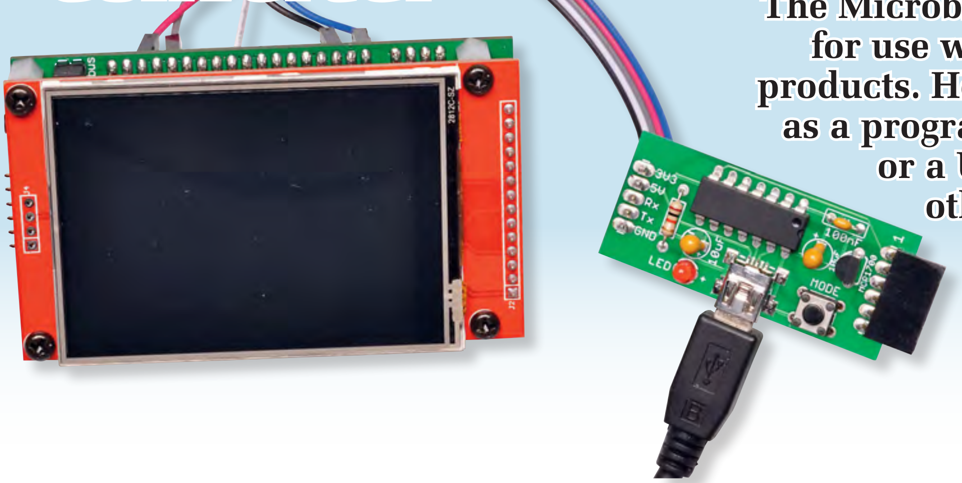
Three of the pins on CON2 (pins 3, 14 and 25) are also connected to the LCD panel for communicating with the display using the SPI serial protocol. For this reason, they cannot be used as general-purpose I/O pins, however, they can still be used by you for SPI communications if needed – this is why they are included on this connector.

The *User Manual* describes how to use the SPI interface simultaneously with the LCD and it is not hard to do. However, for normal operation, you should make sure that you do not use pins 3, 14 and 25 for general I/O.

If you have any issues or questions then contact Phil Boyce via email (**phil.boyce@micromite.org**) and he will be able to assist you. We hope you enjoy using this new version of the *BackPack*.

Microbridge

Cheap universal PIC32 programmer combined with a USB/serial converter



The Microbridge was created as a tool for use with the Micromite range of products. However, it can also be used as a programmer for any PIC32, and/or a USB-to-serial converter for other processors, such as the Arduino or Raspberry Pi.

By Geoff Graham

The Micromite microcontroller, which has featured many times on our pages, requires a USB/serial converter to load, edit and run the program (unless you purchased a pre-programmed chip).

We previously recommended devices based on the CP2102, or FTDI FT232 for this job. They are cheap and convenient; however, you still require a PIC32 programmer if you need to update the Micromite firmware.

Firmware updates for the Micromite are released regularly and usually provide worthwhile new features and bug fixes, so it is definitely an advantage having access to a PIC32 programmer.

But now you don't need a dedicated PIC32 programmer. Instead, the *Microbridge* combines the USB/serial interface and PIC32 programming features in a single package. It is easy to build and uses a low-cost 14-pin chip.

In fact, the Microbridge is so economical and convenient that it makes sense to permanently attach it to your Micromite. With that in mind, we have designed a new version of the *Micromite LCD Backpack* with the *Microbridge* integrated which is featured on page 22 of this issue.

The development of the *Microbridge* and the associated software was truly an international effort, with contributions from New Zealand, Thailand, the US and UK (see the side box for the details).

Circuit details

Referring to Fig.1, you can see that the *Microbridge* consists of just a Microchip PIC16F1455 microcontroller, a voltage regulator and a few passive components.

Microbridge credits

The Microbridge is the result of an international collaboration.

- Peter Mather in the UK wrote the firmware for the PIC16F1455 and the BASIC program for programming a PIC16F1455 using a Micromite (see panel on programming)
- Serge Vakulenko in the USA wrote pic32prog
- Robert Rozee in New Zealand wrote the ASCII ICSP interface for pic32prog
- MicroBlocks (a company in Thailand) developed the original concept of using the PIC16F1455 as both a USB/serial converter and programmer, but did not publish their code for copyright reasons.

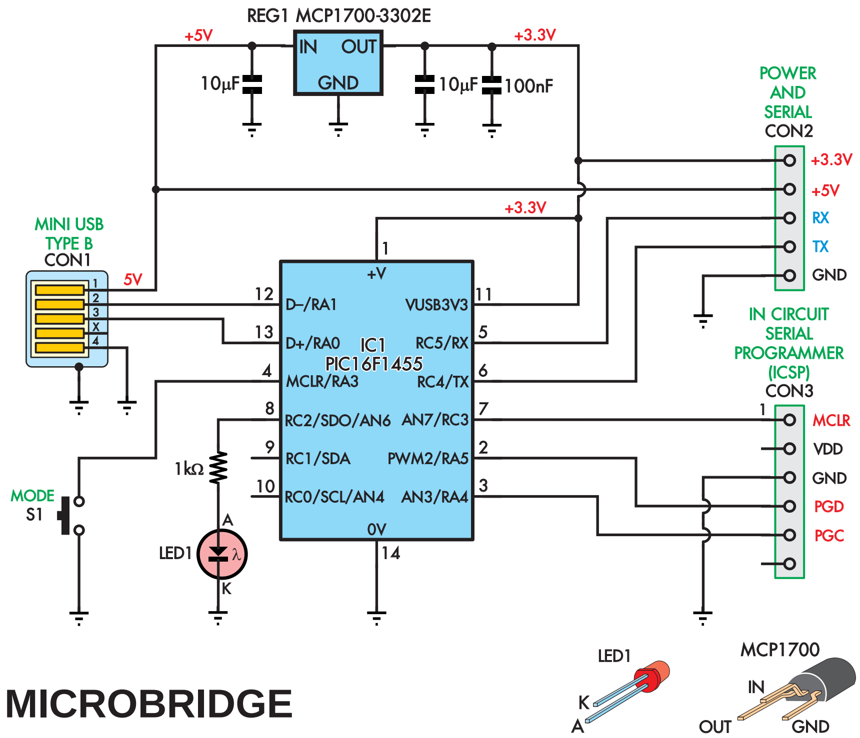
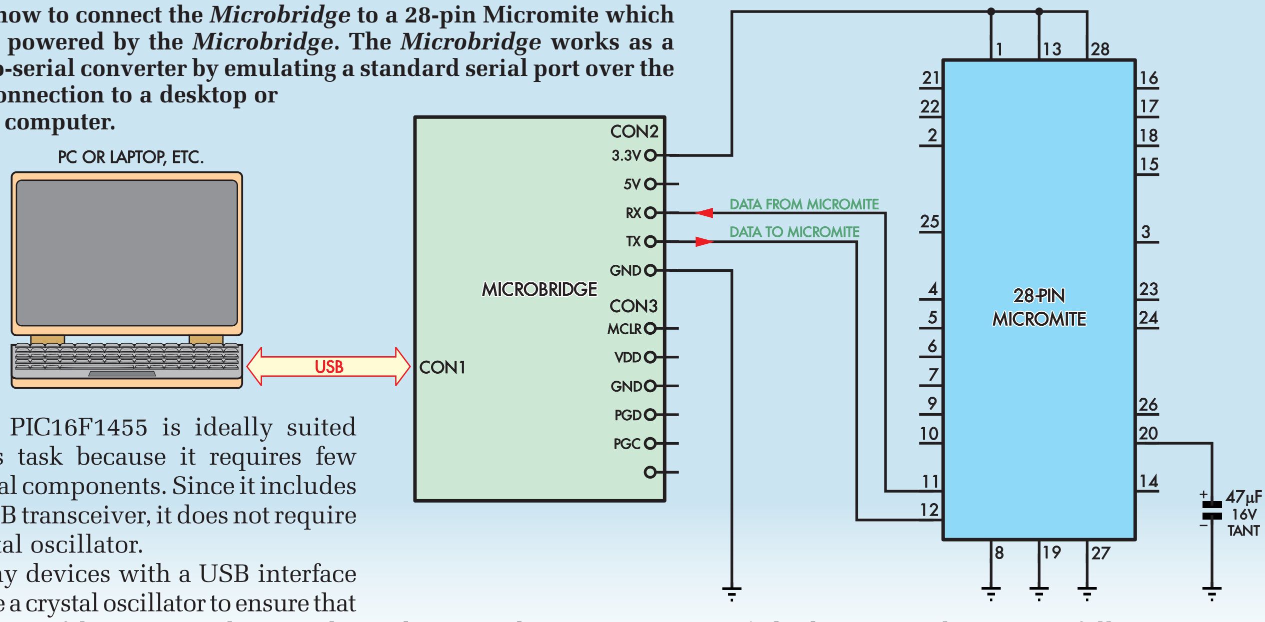


Fig.1: the *Microbridge* consists of a Microchip PIC16F1455 microcontroller, a voltage regulator and a few passive components. The PIC16F1455 is ideally suited to this task because it requires few external components and can automatically tune its internal clock to the host's USB signal timing.

Fig.2: how to connect the *Microbridge* to a 28-pin Micromite which is also powered by the *Microbridge*. The *Microbridge* works as a USB-to-serial converter by emulating a standard serial port over the USB connection to a desktop or laptop computer.



The PIC16F1455 is ideally suited to this task because it requires few external components. Since it includes the USB transceiver, it does not require a crystal oscillator.

Many devices with a USB interface require a crystal oscillator to ensure that the timing of the USB signals meets the strict timing requirements of the USB standard. However, the PIC16F1455 has a feature that Microchip calls 'active clock tuning'.

This allows the PIC16F1455 to use the host's USB signals (which presumably are derived from a crystal oscillator) to automatically tune its internal RC oscillator to the precision required by the standard. Hence, a crystal is not required and this helps keep the circuit simple and the cost down.

The PIC16F1455 can run on a supply voltage of 2.3-5.5V and also includes its own 3.3V regulator for powering its USB transceiver (USB uses 3.3V signal levels).

This means that we could directly power the PIC16F1455 from the USB 5V supply, but then we would need level converters for the signal lines that go to the PIC32 processor (which runs from 3.3V).

For that reason, we've included a low-cost 3.3V regulator (REG1, MCP1700) for powering the PIC16F1455 and we are ignoring its internal regulator. A side benefit of this approach is that this 3.3V supply has spare current capacity so it can also be used to power an attached Micromite chip.

The serial interface is made available on CON2 and includes the 5V USB power and the 3.3V from our on-board regulator.

By default, the serial interface runs at 38400 baud, which is also the default used by the Micromite's console interface.

The programming interface is on CON3 and this provides the standard I/O pins used for In-Circuit Serial Programming (ICSP) on Microchip products. These are as follows:

Pin 1: MCLR/V_{PP} – this is the reset pin for the PIC32 chip and is driven low by the *Microbridge*. It is also used to force the PIC32 into programming mode. On other PICs, this pin is also used as a programming voltage source of around 15V, but the PIC32 generates this internally.

Pin 2: V_{DD} – normally, this is used to detect the power supply voltage for the PIC32, but on the *Microbridge* it is not used.

Pin 3: GND – the ground connection which must go to V_{SS} (ground) on the PIC32.

Pin 4: PGD – the programming data pin, which is bidirectional so that data can be sent to the PIC32 then read back by the *Microbridge's* firmware to verify that programming has been successful and no errors have been introduced.

Pin 5: PGC – the programming clock signal, generated by the *Microbridge* to synchronise the transfer of data on the PGD line.

Pin 6: NC – not connected in most ICSP devices.

The *Microbridge* is switched into programming mode by using pushbutton switch S1. LED1 flashes to indicate serial traffic or it lights up continuously when in programming mode.

Parts List

1 double-sided PCB available from the *EPE PCB Service*, coded 24104171, 50mm × 22.5mm
1 Mini Type-B USB socket, horizontal SMD USB 2.0
1 PCB-mount SPST momentary tactile switch (S1)
1 14-pin DIL IC socket (for IC1)
1 6-pin 90° female socket, 2.54mm pitch OR
1 6-pin female socket, 2.54mm pitch, with pins bent 90°
1 5-pin vertical header, 2.54mm pitch

Semiconductors
1 PIC16F1455-I/P* microcontroller programmed with 2410417A.HEX (IC1)
1 MCP1700-3302E/TO 3.3V linear regulator (REG1)
1 3mm red LED (LED1)

Resistors (5%, ¼W)
1 1kW

Capacitors
2 10µF 16V tantalum or X5R SMD ceramic (3216/1206 size)
1 100nF 50V multi-layer ceramic

* PIC16LF1455-I/P or PIC16(L)F1454-I/P are also suitable

Win a Microbridge!

EPE is running a competition to win a fully-assembled Microbridge thanks to the generous sponsorship of Micromite online shop **micromite.org**

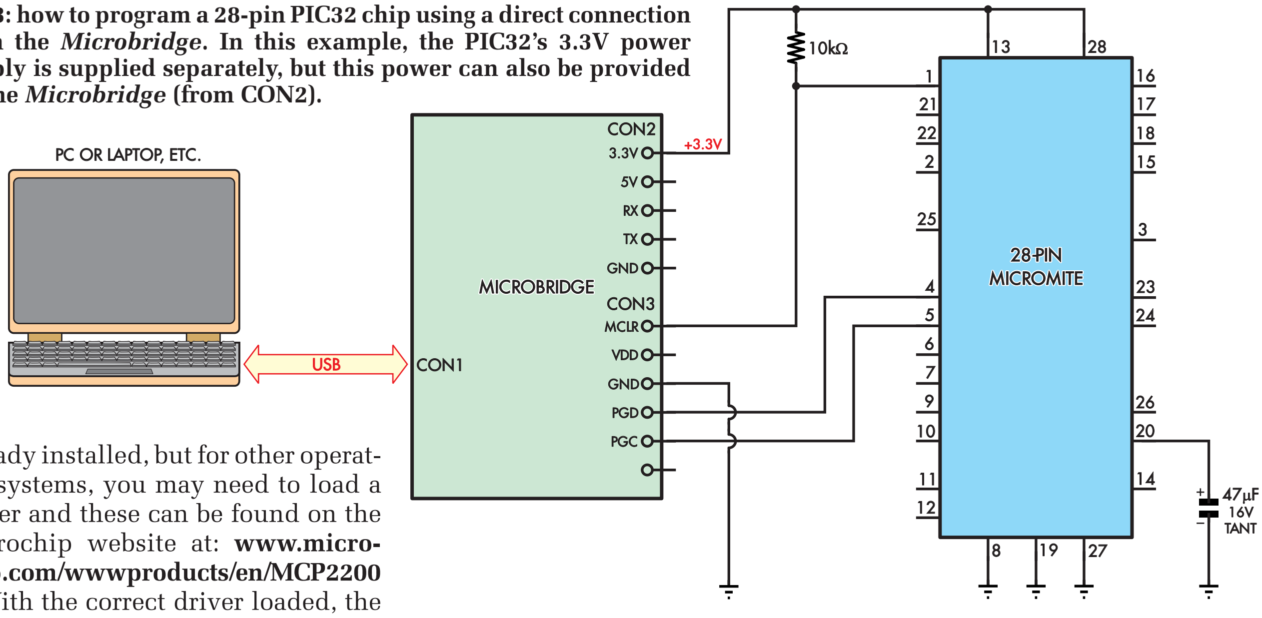
For entry details, please turn to page 27

USB/serial mode

USB/serial mode is the default when power is applied. In this mode, the *Microbridge* works as a USB-to-serial converter – it emulates a standard serial port over USB and converts the signal to a standard TTL-level serial interface for the Micromite (or other processor).

From an operating system viewpoint, the *Microbridge* imitates the Microchip MCP2200 USB/serial converter. Windows 10 is delivered with the correct driver for this device

Fig.3: how to program a 28-pin PIC32 chip using a direct connection from the *Microbridge*. In this example, the PIC32's 3.3V power supply is supplied separately, but this power can also be provided by the *Microbridge* (from CON2).



already installed, but for other operating systems, you may need to load a driver and these can be found on the Microchip website at: **www.microchip.com/wwwproducts/en/MCP2200**

With the correct driver loaded, the *Microbridge* appears as a standard serial port on your computer. For example, in Windows it will appear as COMxx where xx is some number allocated by Windows.

To discover this number you can use Device Manager and look under 'Ports (COM and LPT)' for the *Microbridge*, which will be labelled 'USB Serial Port (COMxx)', where xx is the serial port number (eg, COM6). You can then start your terminal emulator (eg, Tera Term) and specify this COM number in the setup menus.

By default, the *Microbridge* operates at 38400 baud with 8-bit data, one stop bit and no parity, which are the standard settings used by the Micromite's console. However, you can change the baud rate to any standard speed from 300 to 230400 (ie, 300, 600, 1200, 2400, 4800, 9600, 19200, 38400, 57600, 76800, 115200 or 230400 baud) in the terminal emulator.

Fig.2 shows how to connect the *Microbridge* to a 28-pin Micromite, which is also powered by the *Microbridge*. When a character is sent or received by the *Microbridge*, LED1 flashes briefly. This is a handy visual clue that the device is working correctly.

Note that TX (transmit) from the *Microbridge* must go to the RX (receive) on the Micromite; likewise, the TX on the Micromite must connect to RX on the *Microbridge*. This is logical when you think about it because signals transmitted by one device must be received by the other.

If you connect pin 1 of CON3 (the programming connector) to the MCLR (reset) pin of the Micromite, you can also use the *Microbridge* to remotely reset the Micromite. This is done by sending a serial break signal to the *Microbridge*. In Tera Term this is accomplished by pressing ALT-B or via the Tera Term menu.

Another way of generating a reset is to press and hold the mode switch on the *Microbridge* for two or more seconds. LED1 will flash and the MCLR line will be briefly driven low to effect the reset.

Programming mode

CON3 on the *Microbridge* (the ICSP socket) is compatible with the connector used on the Microchip PICkit 3 programmer so the *Microbridge* can plug into any programming connector intended for the PICkit 3. For example, the *Microbridge* can plug directly onto the programming connector on the original *Micromite LCD Backpack* (see the accompanying photograph on the next spread).

Alternatively, to program a 28-pin PIC32 chip using direct connections, Fig.3 shows how to do this. The PIC32's 3.3V power supply can be supplied separately or this power can be provided by the *Microbridge* via CON2.

To enter programming mode, momentarily press and release mode switch S1 and LED1 will illuminate to indicate that programming mode is active.

If you accidently pressed this switch and did not want to enter programming mode, cycle the power on the *Microbridge*, or, press and hold down S1 for two seconds; either way, this will return you to the default USB/serial mode.

To program a PIC32 via the *Microbridge*, use a program called **pic32prog** written by Serge Vakulenko in California.

This is a Windows progra and it can be downloaded from th from GitHub: **https://github.com/se gev/pic32prog**

pic32prog must be run from th command prompt in Windows using th the command line:

pic32prog -d ascii:comxx yyyy.he

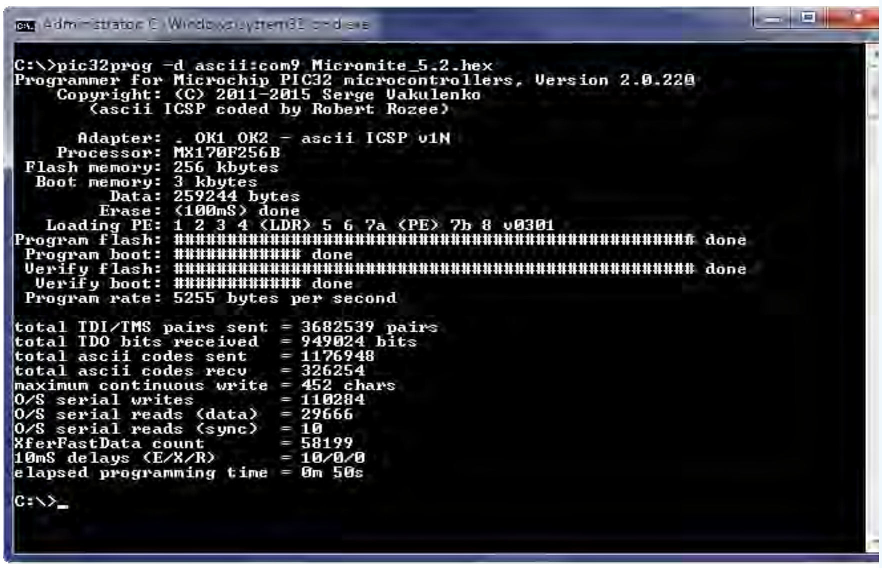


Fig.4: This screenshot shows the complete operation of pic32prog**. It uploads the h file to the *Microbridge*, which programs it into the PIC32 and subsequently reads ba the programmed data to verify that the programming operation completed correct**

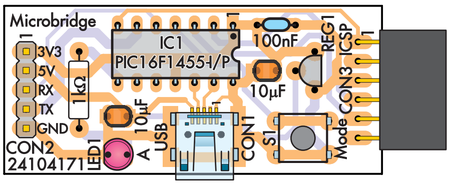


Fig.5: PCB component overlay diagram for the *Microbridge*. The USB socket is the only SMD component. IC1 may be mounted in a socket. We prefer SMD ceramic capacitors to tantalum due to their longer life, however you can use through-hole tantalum capacitors.

Where xx is the COM port number created by Windows for the *Microbridge* and **yyyy.hex** is the file containing the firmware that you want to program into the PIC32. For example, if your *Microbridge* was allocated the virtual serial port of COM12 and the file that you wanted to program was **firm.hex**, the command line that you should use would be:

```
pic32prog -d ascii:com12 firm.hex
```

When you press enter, **pic32prog** will automatically upload the hex file to the *Microbridge*, program it into the PIC32 then read back the programmed data to verify that the programming operation was executed correctly. Fig.4 shows a typical output of this operation.

At the completion of the programming operation, LED1 switches off and the *Microbridge* will revert to operating as a USB/serial converter. You can then start up your terminal emulator, connect to the *Microbridge* and run your program.

A common cause of programming errors is that **pic32prog** cannot access the serial port on your computer because you have not closed the terminal emulator that you were previously using to access the *Microbridge*. So, make sure that you close your terminal emulator before you run **pic32prog**.

Construction

The *Microbridge* uses fewer than a dozen components and all except the USB socket are through-hole types, so construction should take less than half an hour. The component overlay diagram is shown in Fig.5.

Start with the USB socket as this is the only surface-mount component. On the underside of the socket, there should be two small plastic pegs which match corresponding holes on the PCB and these will correctly locate the socket.

Once it is in place, solder the connector's mounting lugs first using plenty of solder for strength then, using a fine point soldering iron tip, solder the signal pins. Carefully check the pin soldering under a good light and with magnification and clean up any solder bridges using solder wick with a little added flux paste to make it easier.

The remaining components are easy to fit and should be soldered starting with the low-profile items such as resistors and ending with the high profile components such as the connectors.

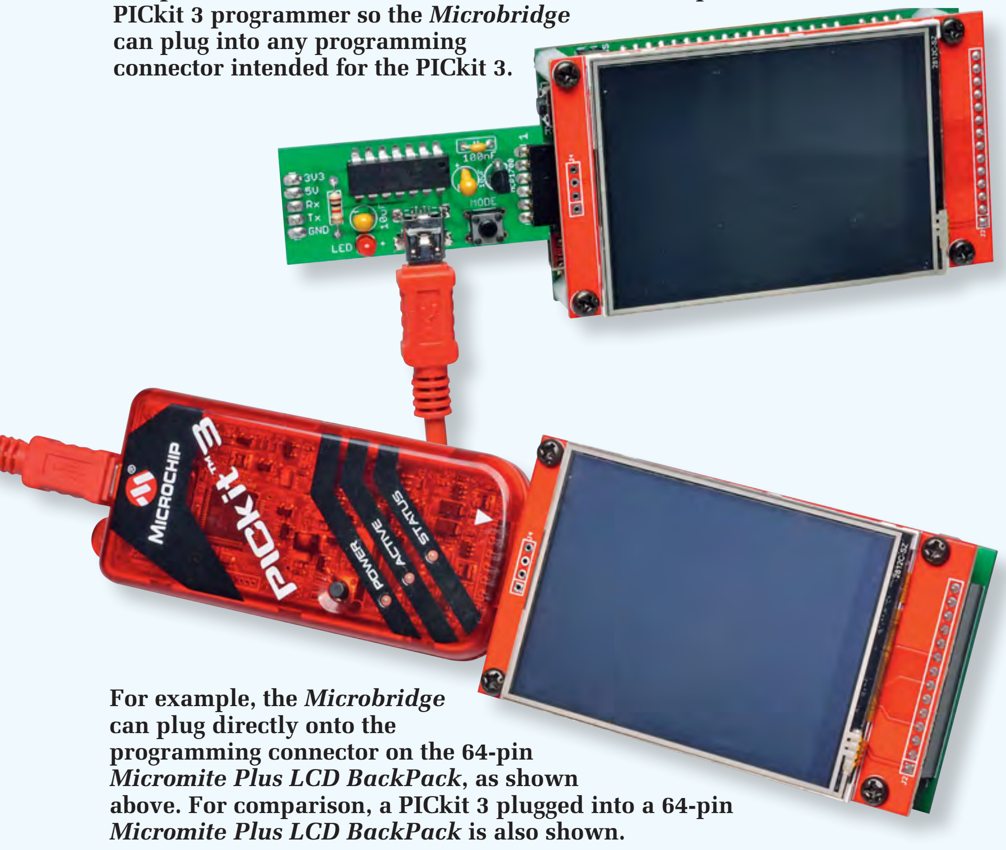
Two of the capacitors and the LED are polarised, so pay attention to their mounting orientation. We did not use an IC socket for IC1 because we had programmed and tested it beforehand, but a socket is recommended and is handy if you suspect a fault and want to swap out the IC for testing.

For CON2 (the serial I/O and power) connector, we mounted a five-pin header on the underside of the board so that it could easily plug into a solderless breadboard for prototyping with the Micromite, but you could use a different arrangement, for example, flying leads.

The right-angle six-pin socket used for the ICSP programmer output (CON3) can be difficult to find so you can do what we did and purchase a straight six-pin socket intended for Arduino boards and bend the pins to 90° so that the socket can mount flush to the PCB. See the parts list for suitable components.

(**Note: although not shown on the circuit diagram, a 10kΩ resistor between PIC pins 1 and 5 will improve S1 response when the Microbridge is not connected to a target device.**)

CON3 on the *Microbridge* (the ICSP socket) is compatible with the connector used on the Microchip PICkit 3 programmer so the *Microbridge* can plug into any programming connector intended for the PICkit 3.



For example, the *Microbridge* can plug directly onto the programming connector on the 64-pin *Micromite Plus LCD BackPack*, as shown above. For comparison, a PICkit 3 plugged into a 64-pin *Micromite Plus LCD BackPack* is also shown.

Testing

There is not much to go wrong with the *Microbridge*, so if it does not work the first time you should first re-check the driver installation on your PC. Do you have the right driver, is it installed correctly and do you have the right COM port number? In normal USB/serial mode the *Microbridge* will draw about 8mA and any reading substantially different from this indicates an assembly error.

A handy test feature is that when you press a key in your terminal emulator, LED1 on the *Microbridge* should flash. Another test that you can make is to short the TX and RX pins on CON2, and as you type characters into the terminal emulator, you should see them echoed back to the terminal emulator.

Programming a blank PIC16F1455

The *Microbridge* uses a PIC16F1455 which acts as a PIC32 programmer to load the firmware into your blank PIC32 microcontroller; for example, to make it into a Micromite. This sounds great because now you do not need a PIC programmer. Or do you?

Firmware transfer

The problem now is getting the *Microbridge's* firmware into the PIC16F1455. One option is to purchase a pre-programmed PIC16F1455 from **micromite.org**. But if you already have at least one Micromite, you can program the PIC16F1455 yourself using just the Micromite and a standard 9V battery.

It is easy to do and will only take 30 seconds. Then, once you have the PIC16F1455 programmed, you can use it to program as many other Micromites as you want!

To get started, wire up the PIC16F1455, the Micromite and the 9V battery as shown in Fig.6.

The best way to do this is on a solderless breadboard or a strip of perforated prototyping board. The battery can be a standard PP3 9V battery and this is used to provide the programming voltage for the PIC16F1455. Only a few milliamps will be drawn from it, and as long as its terminal voltage is 8V or greater it will do the job. The switch used to connect the battery can be as simple as a lead with an alligator clip that can be clipped onto the battery's positive terminal.

The Micromite used for the programming operation can be any version of the Micromite family (ie, a 28-pin Micromite to a 100-pin Micromite Plus) so long as it is running version 5.0 or later of MMBasic. Pins 4 and 5 on the Micromite are used to load the firmware into the PIC16F1455, and all versions have these two pins free.

If for some reason your one does not, you can edit the BASIC program to change the pin assignments (they are defined at the very start of the program).

With everything connected, load the BASIC program **MicrobridgeProg.bas** into the Micromite. This program can be downloaded for free from the author's website (**geoffg.net/microbridge.html**). It will work with all chips that are supported by the *Microbridge* firmware (16F1455, 16F1454, 16LF1454 or 16LF1455). This program was written by Peter Mather of the UK, who also developed the *Microbridge's* firmware.

Make sure that the 9V battery is disconnected and run the BASIC program on the Micromite. From there, it is just a case of following the program's on-screen instructions which will tell you when to connect and disconnect the battery.

The programming time is under 30 seconds and the software will report its progress as it goes. Fig.7 shows a typical programming session. When the programming operation has finished, you can disconnect the battery, remove the PIC16F1455 and install it in your *Microbridge* board. Then, you can use the *Microbridge* to program further PIC32 chips.

The firmware loaded into the PIC16F1455 will be version 1.18 and this contains a bootloader which allows another Micromite to update it via the serial console interface.

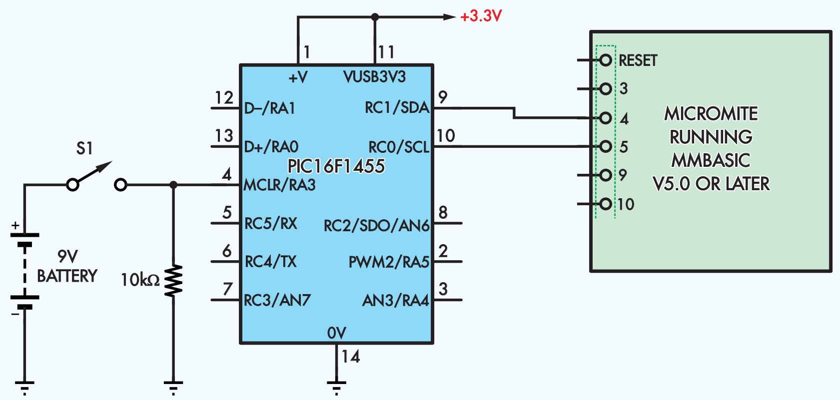


Fig.6: if you already have a Micromite, you can use it to program a blank PIC16F1455 (for use as a *Microbridge*). All you need is a standard 9V battery and a 10kΩ resistor. Connect everything as shown in the circuit above. The MicrobridgeProg.bas **running on the Micromite will prompt you when to connect and disconnect the battery.**

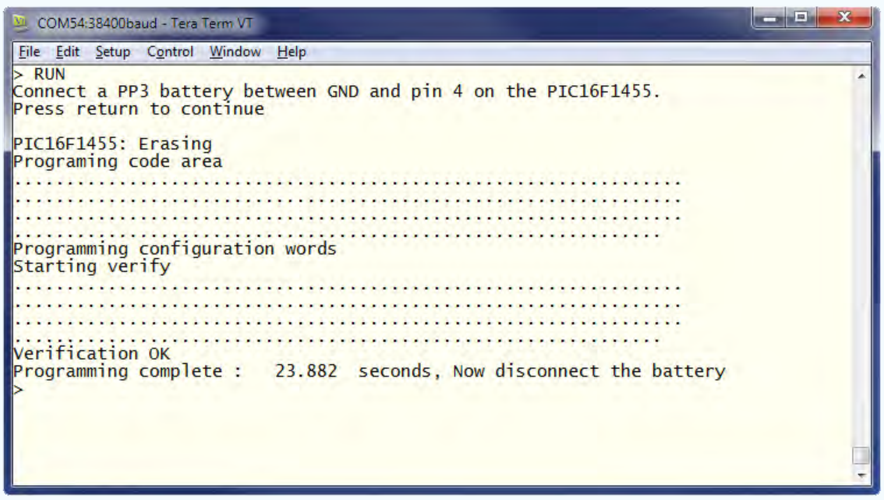


Fig.7: this screenshot shows the complete programming operation for a PIC16F1455 using a Micromite and a standard 9V battery. The program running on the Micromite is MicrobridgeProg.bas**.**

Updates

This updating is even easier than the initial programming described above and can be done with the *Microbridge* permanently connected to the Micromite. There will likely be no need to update the *Microbridge's* firmware but, if there is, the current firmware can do it.

Teach-In 2018

Get testing! – electronic test equipment and measurement techniques

Part 8: Digital measurements

by Mike Tooley

Welcome to *Teach-In 2018: Get testing! – electronic test equipment and measurement techniques*. This *Teach-In* series will provide you with a broad-based introduction to choosing and using a wide range of test gear, how to get the best out of each item and the pitfalls to avoid. We'll provide hints and tips on using, and – just as importantly – interpreting the results that you get. We will be dealing with familiar test gear as well as equipment designed for more specialised applications.

Our previous *Teach-In* series have dealt with specific aspects of electronics, such as PICs (*Teach-In 5*), Analogue Circuit Design (*Teach-In 6*) or popular low-cost microcontrollers (*Teach-In 7* and *8*). The current series is rather different because it has been designed to have the broadest possible appeal and is applicable to all branches of electronics. It crosses the boundaries of analogue and digital electronics with applications that span the full range of electronics – from a single-stage transistor amplifier to the most sophisticated microcontroller system. There really is something for everyone in this series!

Each part includes a simple but useful practical *Test gear project* that will build into a handy gadget that will either extend the features, ranges and usability of an existing item of test equipment or that will serve as a stand-alone instrument. We've kept the cost of these projects as low as possible and most of them can be built for less than £10 (including components, enclosure and circuit board).

This month

In this penultimate part of our current *Teach-In* series, *In theory* will provide an overview of the way in which digital data is represented in a digital system and how it can be captured and analysed. *Get it right!* will help you avoid some potential pitfalls when making digital measurements and will provide useful hints and tips to help you improve the accuracy and relevance of your measurements. Finally, our *Test gear project* features a handy logic probe that can be used with a wide range of different types of digital logic.

In theory: Digital signals and logic levels

This month, we turn our attention to the tests and measurements that need to be made in a digital rather than analogue world. Since we are only dealing with two logical states – variously referred to as 'on' or 'off', 'high' or 'low', and 'logic 1' or 'logic 0' – this should be quite easy. In practice, however, this can be rather different since the logical states that we need to view are often rapidly changing and may need to be captured so that we can view and make sense of them. Even more challenging is that we often need to capture data on multiple signal lines so that we can view and analyse the relationship between signals within the time domain. This calls for some sophisticated but not necessarily expensive items of test equipment. We will start by looking at how digital signals are represented before moving on to describe ways in which they can be measured and analysed.

Representing digital signals

Logic levels

Logic levels are simply the range of voltages used to represent the logical states 0 and 1. The logic levels for CMOS differ markedly from those associated with TTL. CMOS devices are usually able to operate over a wide range of supply voltage (from 3V to 15V for most standard devices) and their logic levels are relative to the supply voltage used (with roughly one third and two thirds of the supply voltage marking the upper boundaries of logic 0 and the lower boundary of logic 1 respectively). By contrast, the logic levels associated with standard TTL devices tend to be more precise and absolute. Table 8.1 shows the normally accepted range of values for conventional 5V CMOS, TTL, and low-voltage (LV) TTL/CMOS devices.

Noise margin

Logical states need to stand out against other signals that may be present in an electronic circuit. In other words, a logic 1 should unambiguously be a logic 1. Likewise, a logic 0 should definitely be a logic 0. Of concern here is the need for sufficient separation between the two defined states. In other words, there should be a clear boundary between them. The difference between the logic 0 and logic 1 boundaries is known as the *noise margin*; it is an important parameter associated with any logic family. Put simply, noise margin is a measure of the ability of the device to reject noise; the larger the noise margin the better its ability to perform in an environment in which noise is present (and where the superimposed noise may be sufficient to cross the logic level boundaries).

Noise margin is usually defined in terms of the difference between the minimum

Table 8.1 Logic levels

Logic state	CMOS 5V	TTL 5V	LV TTL/CMOS 3.3V
Logic 1 (high)	>3.5V	>2V	>2V
Logic 0 (low)	<1.5V	< 0.8V	<0.8V
Threshold	2.5V	1.5V	1.5V
Indeterminate region	1.5V to 3.5V	0.8V to 2V	0.8V to 2V

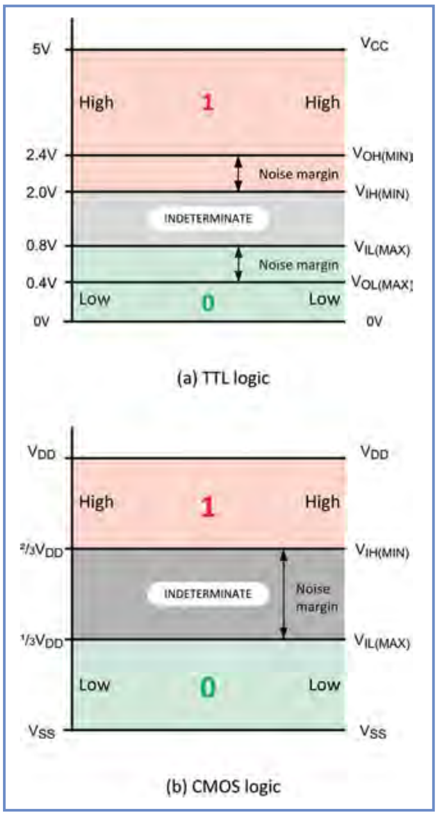


Fig.8.1. Comparison of logic levels and noise margins for standard 5V TTL and CMOS devices

values of high state output and high state input voltage and the maximum values of low state output and low state input voltage. Hence:

Noise margin = $V_{OH(MIN)} - V_{IH(MIN)}$

or

Noise margin = $V_{OL(MAX)} - V_{IL(MAX)}$

Where $V_{OH(MIN)}$ is the minimum value of high-state (logic 1) output voltage, $V_{IH(MIN)}$ is the minimum value of high-state (logic 1) input voltage, $V_{OL(MAX)}$ is the maximum value of low-state (logic 0) output voltage, and $V_{IL(MAX)}$is the minimum value of low-state (logic 0) input voltage. The noise margin for the legacy 7400 TTL series is typically 400mV, while for 5V CMOS it is approximately 2V, as illustrated in Fig.8.1.

In practice, you are likely to encounter a variety of sub-families of the original 'standard' TTL and CMOS logic families. These include CMOS devices compatible with TTL (HCT and FCT) as well as low-voltage (LV) logic devices. The chart shown in Fig.8.2 provides a useful comparison of the threshold voltages of these different families.

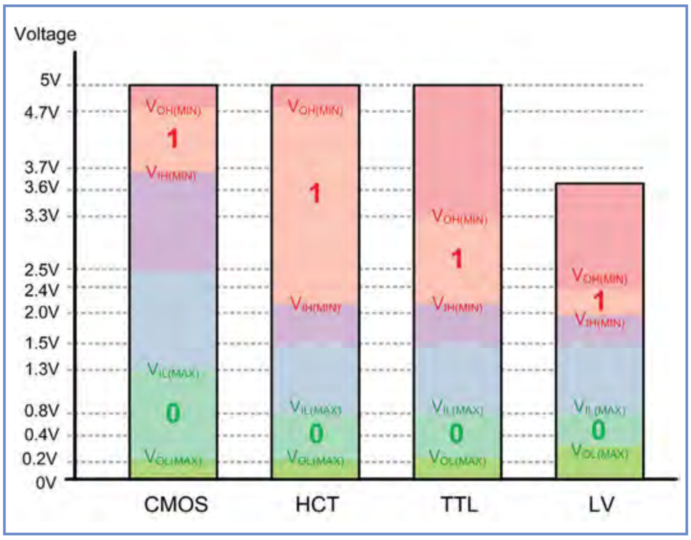


Fig.8.2. Comparison chart of common logic families

Logic gates

Basic logical operations (eg, AND, OR) are carried out by means of individual circuits known as 'gates'. The symbols for some basic logic gates are shown, together with their truth tables in Fig.8.3. The action of each of the basic logic gates is summarised below. Note that while inverters and buffers each have only one input, exclusive-OR and exclusive-NOR gates have two inputs and the other basic gates (AND, OR, NAND and NOR) are commonly available with up to eight inputs (but for these there is no theoretical limit).

Buffers

Buffers do not affect the logical state of a digital signal (ie, a logic 1 input results in a logic 1 output and a logic 0 input results in a logic 0 output). Buffers are normally used to provide extra current drive at the output but can also be used to regularise the logic levels present at an interface.

Inverters

Inverters are used to complement the logical state (ie, a logic 1 input results in a logic 0 output and vice versa). Inverters also provide extra current drive and, like buffers, are used in interfacing applications where they provide a means of regularising logic levels present at the input or output of an LSI device.

AND gates

These gates will only produce a logic 1 output when all inputs are simultaneously at logic 1. Any other input combination results in a logic 0 output.

OR gates

These gates will produce a logic 1 output whenever one or more of their inputs are at logic 1. Putting this another way, an OR gate will only produce a logic 0 output whenever all inputs are simultaneously at logic 0.

NAND gates

These gates will only produce a logic 0 output when all inputs are simultaneously at logic 1. Any other input combination will produce a logic 1 output. A NAND gate, therefore, is nothing more than an AND gate with its output inverted. The circle shown at the output denotes this inversion.

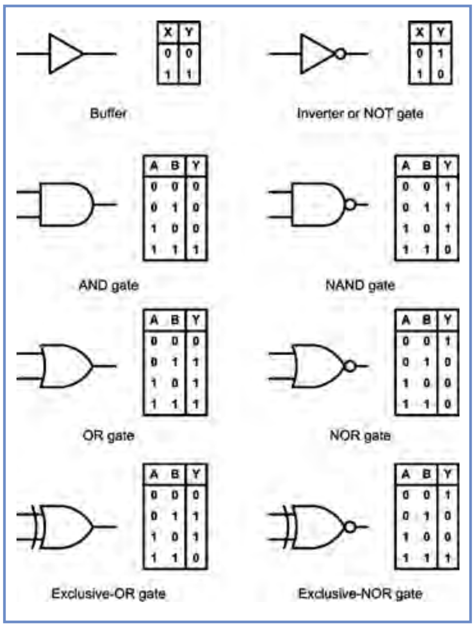


Fig.8.3. Logic gate symbols and truth tables

NOR gates

These gates will only produce a logic 1 output when all inputs are simultaneously at logic 0. Any other input combination will produce a logic 0 output. A NOR gate, therefore, is simply an OR gate with its output inverted. A circle is again used to indicate inversion.

Exclusive-OR gates

Exclusive-OR gates Sometimes written as 'XOR') will produce a logic 1 output whenever either one of the inputs is at logic 1 and the other is at logic 0. Exclusive-OR gates produce a logic 0 output whenever both inputs have the same logical state (ie, when both are at logic 0 or both are at logic 1).

Monostable

A logic device which has only one stable output state is known as a 'monostable'. The output of such a device is initially at logic 0 (low) until an appropriate level change occurs at its trigger input. This level change can be from 0 to 1 (positive-edge trigger) or 1 to 0 (negative-edge trigger) depending upon the particular monostable device or configuration. Upon receipt of a valid trigger pulse the output of the monostable changes state to logic 1. Then, after a time interval determined by external C-R timing components, the output reverts to logic 0. The device then awaits the arrival of the next trigger. A typical application for a monostable device is in stretching a pulse of very short duration.

Bistables

The output of a bistable can take one of two stables states, either logic 0 or

Table 8.2 Characteristics of logic levels

Characteristic	Logic family			
	74	74LS	74HC	40BE
Maximum supply voltage	5.25V	5.25V	5.5V	18V
Minimum supply voltage	4.75V	4.75V	4.5V	3V
Static power dissipation (mW per gate at 100kHz)	10	2	negligible	negligible
Dynamic power dissipation (mW per gate at 100kHz)	10	2	0.2	0.1
Typical propagation delay (ns)	10	10	10	105
Maximum clock frequency (MHz)	35	40	40	12
Speed-power product (pJ at 100kHz)	100	20	1.2	11
Minimum output current (mA at VOUT = 0.4V)	16	8	4	1.6
Fan-out (LS loads)	40	20	10	4
Maximum input current (mA at VIN = 0.4V)	–1.6	–0.4	0.001	–0.001

logic 1. Once *set*, the output of a bistable will remain at logic 1 for an indefinite period until the bistable is *reset*, at which time the output will revert to logic 0. A bistable thus constitutes a simple form of *memory cell* because it will remain in its latched state (either *set* or *reset*) until commanded to change its state (or until the supply is disconnected). Popular forms of bistable include R-S, D and J-K types.

R-S bistables

The simplest form of bistable is the R-S bistable. This device has two inputs, SET and RESET, and complementary outputs, Q and $\overline{Q}$. A logic 1 applied to the SET input will cause the Q output to become (or remain at) logic 1 while a logic 1 applied to the RESET input will cause the Q output to become (or remain at) logic 0. In either case, the bistable will remain in its SET or RESET state until an input is applied in such a sense as to change the state.

D-type bistables

The D-type bistable has two principal inputs; D (standing variously for data or delay) and CLOCK (CK). The data input (logic 0 or logic 1) is clocked into the bistable such that the output state only changes when the clock changes state. Operation is thus said to be *synchronous*. Additional subsidiary inputs (which are invariably active low) are provided, which can be used to directly set or reset the bistable. These are usually called PRESET (PR) and CLEAR (CLR). D-type bistables are commonly used as data latches (a simple form of memory) and as binary dividers.

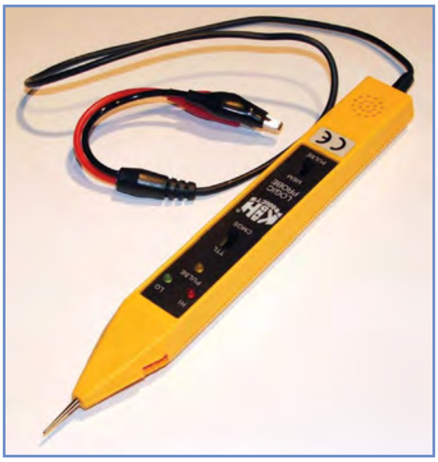


Fig.8.4. A typical logic probe

J-K bistables

J-K bistables are the most sophisticated and flexible of the bistable types, and they can be configured in various ways including binary dividers, shift registers, and latches. J-K bistables have two clocked inputs (J and K), two direct inputs (PRESET and CLEAR), a CLOCK (CK) input, and outputs (Q and $\overline{Q}$). As with R-S bistables, the two outputs are complementary (ie, when one is 0 the other is 1, and vice versa). Similarly, the PRESET and CLEAR inputs are invariably both active low (ie, a 0 on the $\overline{\text{PRESET}}$ input will set the Q output to 1 whereas a 0 on the $\overline{\text{CLEAR}}$ input will set the Q output to 0).

Logic gate characteristics

Table 8.2 summarises the key characteristics of the original members of the TTL family with the equivalent CMOS logic. There are some important points worth noting:

- CMOS devices are static sensitive and require appropriate anti-static handling techniques
- CMOS logic operates over a much larger range of supply voltage than conventional TTL
- CMOS devices tend to be much slower than their TTL counterparts
- TTL devices consume significantly more power than their CMOS counterparts
- TTL devices are capable of driving more loads than CMOS devices
- CMOS devices require negligible input current and impose minimal load on an input.

Logic probes

The simplest and most convenient method of examining logic states involves the use of a logic probe. When making measurements on digital circuits, this handy gadget is much easier to use than a digital multimeter or an analogue oscilloscope. It comprises a hand-held probe fitted with LEDs that indicate the logical state of its probe tip.

Unlike a digital multimeter, a logic probe can usually distinguish between lines which are actively pulsing, and those that are in a permanently *tri-state* (effectively disconnected) condition. In the case of a line which is being pulsed, the logic 0 and logic 1 indicators will both be illuminated (though not necessarily with the same brightness) whereas, in the case of a tri-state line neither indicator should be constantly illuminated.

Logic probes generally also provide a means of displaying pulses having a very short duration, which may otherwise go undetected. A *pulse stretching* circuit is usually incorporated within the probe circuitry so that an input pulse of very short duration is elongated sufficiently to produce a visible indication on a separate pulse LED.

Logic probes invariably derive their power supply from the circuit under test and are connected by means of a short length of twin flex fitted with insulated crocodile clips (see Fig.8.4). Note that it is essential to ensure that the supply voltage is the same as that used to supply the logic devices on test.

A typical logic probe circuit suitable for home construction is shown in Fig.8.5. This circuit uses a dual comparator to sense the logic 0 and logic 1 levels and a timer, which acts as a monostable pulse stretcher to indicate the presence of a pulse input rather than a continuous logic 0 or logic 1 condition. Typical logic probe indications and waveforms are shown in Fig.8.6.

Fig.8.7 shows how a logic probe can be used to check a simple arrangement of logic gates. The probe is moved from node to node and the logic level is displayed and compared with the expected level. Fig.8.8 shows how a logic probe can be used to test a much more complex circuit in the shape of a modern Mini-ITX computer system.

Logic pulsers

It is sometimes necessary to simulate the logic levels generated by a peripheral

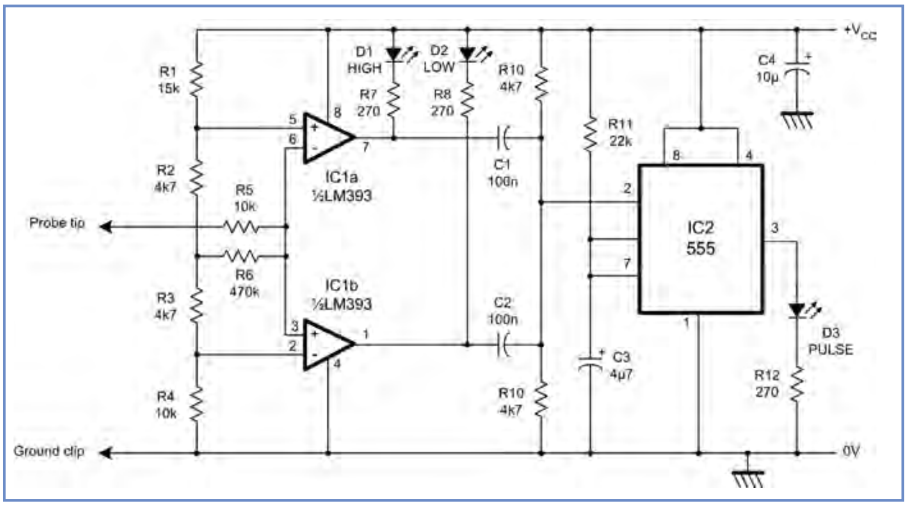


Fig.8.5. A logic probe circuit suitable for home construction

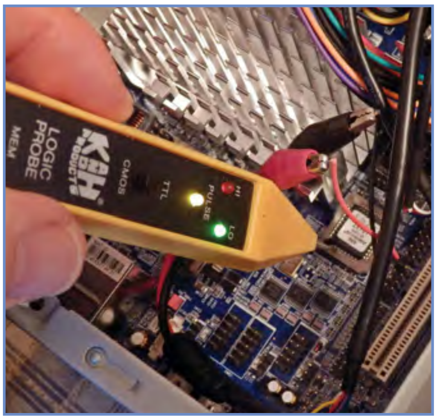


Fig.8.8. Using a logic probe to check the signals present on the BIOS chip of a Mini-ITX motherboard. The probe indicates a signal that is mostly low but also pulsing high (see Fig.8.6)

device or sensor. A permanent logic level can easily be generated by pulling a line up to the logic supply by means of 1kΩ resistor or by temporarily tying a line down to 0V. However, on other occasions, it may be necessary to simulate a pulse rather than a permanent logic state and this can be achieved by means of a hand-held logic pulser.

LED INDICATOR			STATE INDICATED	WAVEFORM
LOW	PULSE	HIGH		
OFF	OFF	ON	Steady logic 1	
ON	OFF	OFF	Steady logic 0	
OFF	OFF	OFF	Open circuit or undefined level	
OFF	BLINK	OFF	Pulse train of near 50% duty cycle at >1MHz	
ON	BLINK	ON	Pulse train of near 50% duty cycle at <1MHz	
OFF	BLINK	ON	Pulse train of high mark:space ratio	
ON	BLINK	OFF	Pulse train of low mark:space ratio	

Fig.8.6. Typical logic probe indications

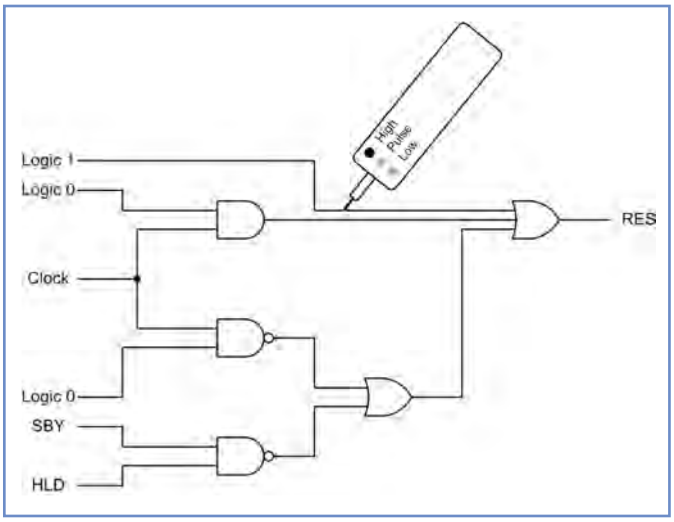


Fig.8.7. Using a logic probe to check a basic logic gate arrangement

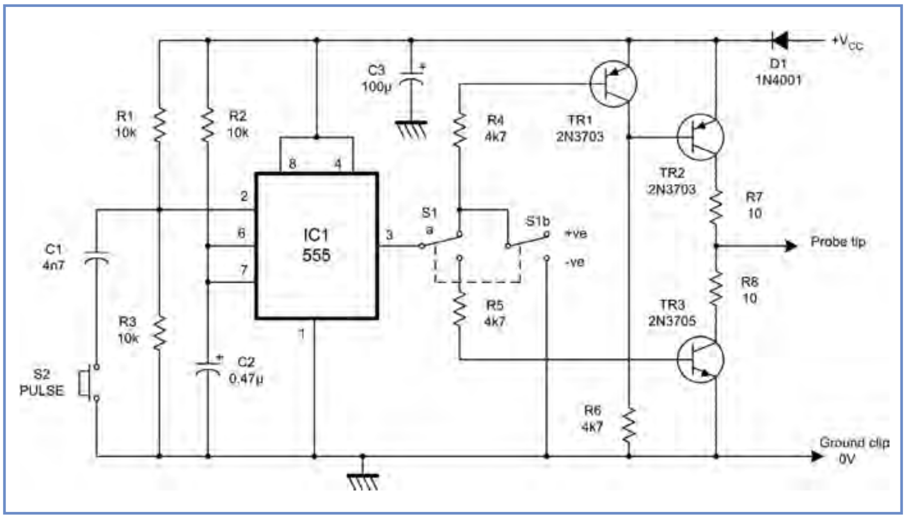


Fig.8.9. A simple logic pulser suitable for home construction

A logic pulser provides a means of momentarily forcing a logic level transition into a circuit regardless of its current state and thus overcomes the need to disconnect or de-solder any of the devices. The polarity of the pulse (produced at the touch of a button) is adjusted so that the node under investigation is momentarily forced into the opposite logical state. During the period before the button is depressed and for the period after the pulse has been completed, the probe tip adopts a tri-state (high impedance) condition. Hence the probe does not permanently affect the logical state of the point in question.

Logic pulsers derive their power supply from the circuit under test in the same manner as logic probes. Here again, it is essential to use the correct logic supply voltage.

A typical logic pulser circuit is shown in Fig.8.9. The circuit comprises a 555 monostable pulse generator triggered from a push-button. The output of the pulse generator is fed to a complementary transistor arrangement in order to make it fully TTL-compatible. As with the logic pulser, this circuit derives its power from the circuit under test.

Fig.8.10 shows an example of the combined use of a logic pulser and a logic probe for testing a simple J-K bistable. The logic probe is used to check the initial state of the Q and $\overline{Q}$ outputs of the bistable, as shown in Fig.8.10 (a) and (b). Note that the Q and $\overline{Q}$ outputs should be complementary. Next, the logic pulser is applied to the clock (CK) input of the bistable (see Fig.8.10(c)) and the Q output is checked using the logic probe. The application of a pulse (using the trigger button) should cause the Q output of the bistable to change state (see Fig.8.10 (d)).

Serial data communication

With anything more than the most basic logic application there's a need for digital data to be exchanged between participating devices. For example, a microcontroller, LCD display and a wide variety of sensors can all be linked

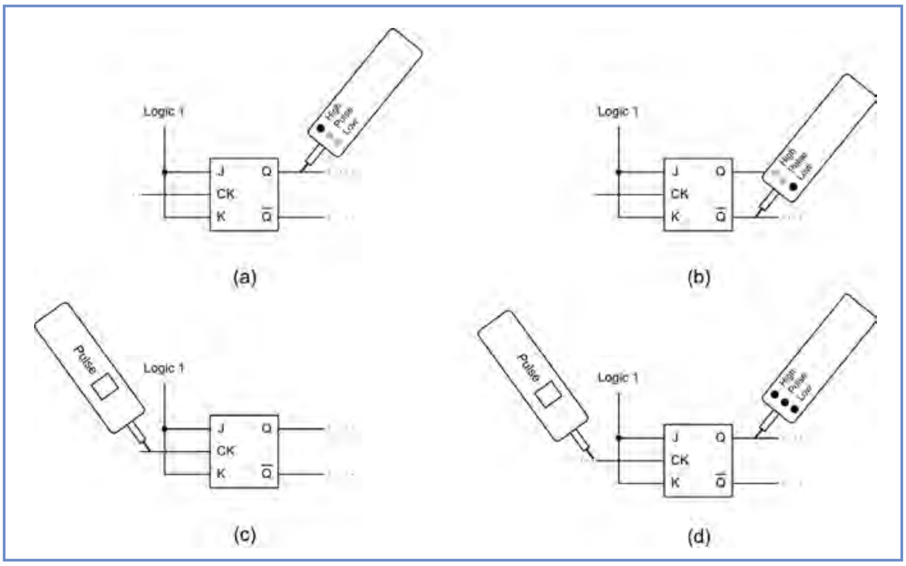


Fig.8.10. A simple logic pulser suitable for home construction

together using one of the popular serial bus connections based on one or more of today's popular and universally available standards such as RS-232, SPI, I^{2}C, and USB).

Serial data communication involves sending a stream of bits, one after another, along a transmission path. Since the data present on a microprocessor bus exists primarily in parallel form, serial I/O techniques are somewhat more complex than those used for simple parallel input and output; serial input data must be converted to parallel (byte wide) data in a form which can be presented to the bus. Conversely, serial output data must be produced from the parallel data present on the internal data bus.

Serial data may be transferred in either synchronous or asynchronous mode. In the former case, transfers are carried out in accordance with a common clock signal (the clock must be available at both ends of the transmission path). Asynchronous operation, on the other hand, involves transmission of data in small packets; each packet containing the necessary information required to decode the data that it contains. Clearly this technique is more complex, but it has the considerable advantage that a commonly available clock signal is not required.

As with programmable parallel I/O devices, a variety of different names are used to describe programmable serial I/O devices, but the asynchronous communications interface adaptor (ACIA) and universal asynchronous receiver/transmitter (UART) are both commonly encountered in serial data communications. Signal connections commonly used with serial I/O devices include:

- *Dn:* Data input/output lines to/from the internal bus
- *RXD:* Received Data (incoming serial data)
- *TXD:* Transmitted Data (outgoing serial data)
- *CTS:* Clear To Send. This (usually active low) signal is taken low by the peripheral when it is ready to accept data from the microprocessor system
- *RTS:* Request To Send. This (usually active low) signal is taken low by the microprocessor system when it is about to send data to the peripheral.

With simple systems (including most popular microcontrollers) signals from serial I/O devices are invariably logic compatible. It should be noted that in general, such signals are unsuitable for anything other than short distance transmission. Reliable data transmission over a greater distance may require specialised line drivers and receivers to provide buffering and level shifting. In noisy environments it can also be advantageous to use balanced transmission using differential signals.

The RS-232D interface is a well-established standard for serial communication between microcontrollers and a wide range of other devices. The original standard dates back to 1987 and is in accordance with international standards CCITT V24, V28 and ISO IS2110. One notable advantage of the RS-232D standard is that it incorporates facilities for loop-back testing, in which data can be sent back to an originating device by looping the TXD line back to the RXD line (see later).

- *Data (TXD, RXD):* RS-232 provides for two independent serial data channels (described as primary and secondary). Both channels provide for full duplex operation (ie, simultaneous transmission and reception). Note that, in practice, both channels are often not used.
- *Handshake control (RTS, CTS):* handshake signals provide the means by which the flow of serial data is controlled, allowing, for example, a DTE to open a dialogue with the DCE prior to actually transmitting data over the serial data path.
- *Timing (TC, RC):* for synchronous (rather than the more usual asynchronous) mode of operation, it is necessary to pass clock signals between the devices. These timing signals provide a means of synchronising the received signal to allow successful decoding.

In practice, few RS-232 implementations make use of the secondary channel featured in the original specification and, since asynchronous (non-clocked) operation is almost invariably used with microcomputer systems, only eight or nine of the original 25 signal lines are regularly used. These lines have the functions shown in Table 8.3.

In asynchronous RS-232 systems, data is transmitted asynchronously as a series of small packets. Each packet represents a single ASCII (or control) character and it must contain sufficient information for the packet to be decoded without the need for a separate clock signal.

ASCII characters are represented by seven bits. The upper-case letter 'A', for example, is represented by the seven-bit binary word 1000001. To send the letter 'A' via RS-232 extra bits must be added to indicate the start and end of the data packet. These are known as the start and stop bits respectively. In addition, we may wish to include a further bit to provide a simple parity-error detecting facility.

Let's look at an example where there is one start bit, seven data bits, one parity bit and two stop bits. The start of the data packet is signaled by the start bit, which is always low irrespective of the contents of the packet. The seven data bits representing the ASCII character follow the start bit. A parity bit is added to make the resulting number of 1s in the group either odd (odd parity) or even (even parity). Finally, two stop bits are added. These are both high. The TTL representation of this character is shown in Fig.8.11 (a).

Table 8.3 Nine-pin RS-232 configuration

Pin	Designation	Function
1	FG	Ground connection
2	TXD	Serial Transmitted data
3	RXD	Serial Received data
4	DTR or RTS	Data Terminal Ready or Request To Send
5	CTS	Clear To Send
6	DSR	Data Set Ready
7	SG	Signal Ground
8	DTR	Data Terminal Ready
9	RI	Ring Indicator

Important note: Not all signals are implemented with current equipment and some pins may be used for different functions. For example, pin-9 is sometimes used for a positive logic supply voltage.

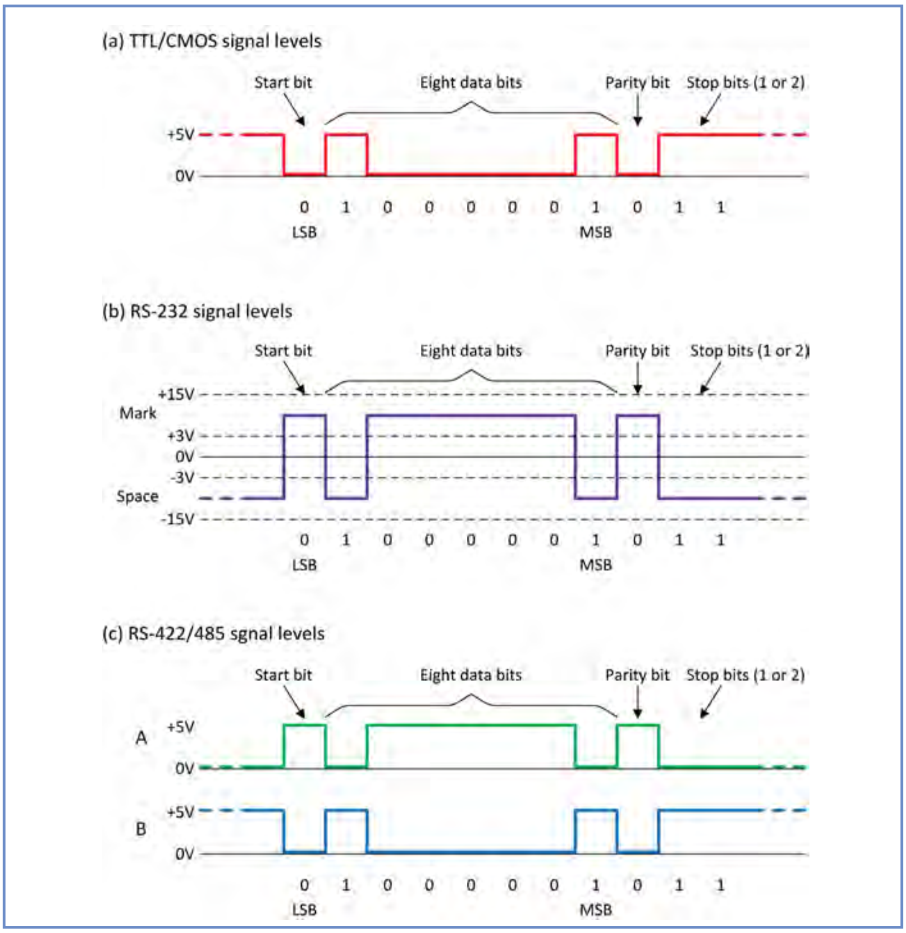


Fig.8.11. Serial data representation

The complete asynchronously transmitted data word thus comprises eleven bits (note that only seven of these actually contain data). In binary terms the word can be represented as: 01000001011. In this example, even parity has been used and thus the ninth (parity bit) is a 0. One of the most commonly used RS-232 schemes involves eight data bits, no parity bit and one stop bit. This is commonly referred to as '8N1'.

The voltage levels employed in a true RS-232 data interface are markedly different from those used within a microcomputer system. A positive voltage (of between +3V and +25V) is used to represent a logic 0 (or SPACE) while a negative voltage (of between –3V and –25V) is used to represent a logic 1 (or MARK). The line signal corresponding to the ASCII character 'A' is shown in Fig.8.11 (b). The level shifting (from TTL to RS-232 signal levels, and vice versa) is usually accomplished using line drivers and line receivers.

Other standards

To overcome some of the limitations of the original RS-232 specification several further standards have been introduced. These generally provide for better line matching, increased distance capability and faster data rates. Notable among these systems are RS-422 (a balanced system which caters for a line impedance as low as 50Ω), RS-423 (an unbalanced system which will tolerate a line impedance of 450Ω minimum), and RS-449 (a very fast serial data standard that uses several modified circuit functions and a 37-way D connector).

RS-422

The RS-422 interface is a balanced system (differential signal lines are used) that employs lower line voltage levels than those used with RS-232. SPACE is represented by a line-voltage level in the range +2V to +6V, while MARK is represented by a line-voltage level in the range, 2V to –6V (see Fig.8.11). RS-422 caters for a line impedance of as low as 50Ω and supports data rates up to 10Mbps.

RS-423

Unlike RS-422, RS-423 employs an unbalanced line configuration (a single signal line is used in conjunction with signal ground). Line voltage levels of +4V to +6V and –4V to –6V represent SPACE and MARK respectively and the standard specifies a minimum line terminating resistance of 450Ω. RS-423 supports a maximum data rate of 100kbps.

RS-449

The RS-449 interface is a further enhancement of RS-422 and RS-423; it caters for data rates up to 2Mbps and provides for upward compatibility with RS-232. Ten extra circuit functions have been provided, while three of the original interchange circuits have been abandoned. In order to minimise confusion, and since certain changes have been made to the definition of circuit functions, a completely new set of circuit abbreviations has been developed. In addition, the standard requires 37-way and 9-way D-connectors, the latter being necessary where use is made of the secondary channel interchange circuits.

Data communication test equipment

Several specialised test instruments and accessories are available for testing data communication equipment, including the following items.

Patch boxes

These low-cost devices facilitate the cross connection of RS-232 (or equivalent) signal lines. The equipment is usually fitted with two D-type connectors (or ribbon cables fitted with a plug and socket) and all lines are brought out to a patching area into which links may be plugged. In use, these devices are connected in series

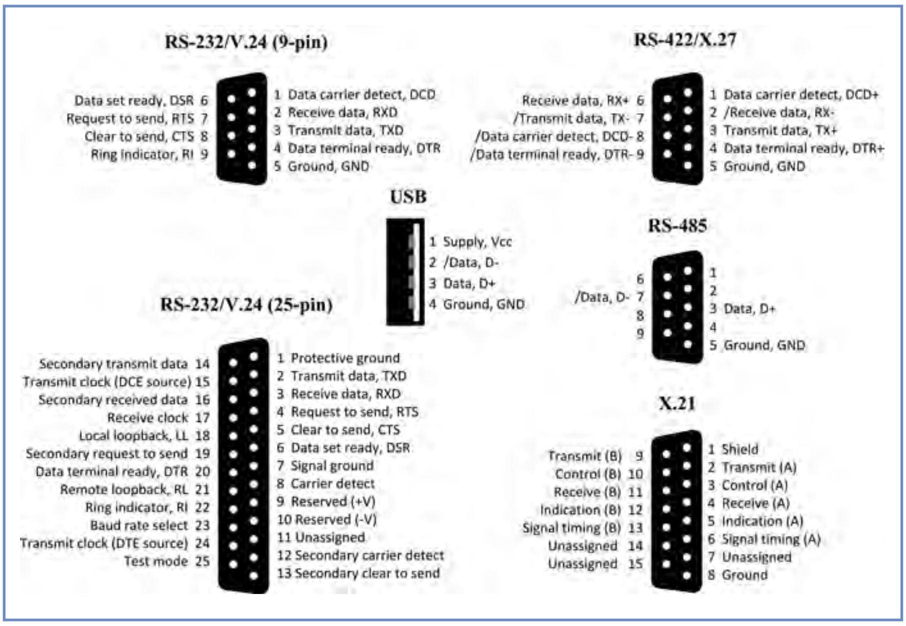


Fig.8.12. Pin connections used for some popular data communication interfaces

with the RS-232 serial data path and various patching combinations are tested until a functional interface is established. If desired, a dedicated cable may then be manufactured in order to replace the patch box.

Gender changers
These normally comprise an extended RS-232 connector that has a male connector at one end and a female connector at the other. Gender changers permit mixing of male and female connector types (note that the convention is male at the DTE and female at the DCE).

Null modems
Like gender changers, these devices are connected in series with an RS-232C serial data path. Their function is simply that of changing the signal lines so that a DTE is effectively configured as a DCE. Null modems can easily be set up using a patch box or manufactured in the form of a dedicated null-modem cable.

Line monitors
These display the logical state (in terms of MARK or SPACE) present on the most commonly used data and handshaking signal lines. LEDs provide the user with a rapid indication of which signals are present and active within the system.

Breakout boxes
These provide access to the signal lines and invariably combine the features of patch box and line monitor. In addition, switches or jumpers are usually provided for linking lines on either side of the box. Connection is almost invariably via two 25-way ribbon cables terminated with connectors.

Oscilloscopes
An oscilloscope can be used to display waveforms of signals present on data lines. It is thus possible to detect the presence of noise and glitches as well as measuring signal voltage levels and rise and fall times. A compensated (×10) oscilloscope probe will normally be required in order to minimise distortion caused by test-lead reactance. A digital storage facility can be invaluable when displaying transitory data.

Interface testers
These are somewhat more complex than simple breakout boxes and generally incorporate facilities for forcing lines into MARK or SPACE states, detecting ‘glitches’, measuring baud rates, and also displaying the format of data words. Such instruments are, not surprisingly, rather expensive.

Multimeters
A general-purpose multimeter (see Part 1) can be useful when testing static line voltages, cable continuity and terminating resistances. A standard multi-range digital instrument will be adequate for most applications, and an audible continuity testing range can be useful when checking data cables.

Universal Serial Bus (USB)

Offering true plug-and-play capability coupled with high data rates, the universal serial bus (USB) has become the de-facto standard for interconnecting a wide range of microcontrollers and computers to an equally wide range of peripheral devices, sharing the available bandwidth through a host-scheduled, token-based protocol. In a conventional USB connection, the USB data (D+ and D–) and power (V_{BUS} and GND) are carried using a four-wire shielded cable. V_{BUS} is nominally +5V at the source, and cable lengths can be up to several metres. To guarantee input voltage levels and proper termination impedance, biased terminations are normally used at each end of the cable.

One of the advantages of USB over other bus systems is its ability to support hot-connection and hot-disconnection from the bus. This important feature requires that the host’s system software is not only able to recognise the connection and disconnection of devices, but is also able to reconfigure the system dynamically. All modern operating systems have this facility.

USB devices attach to the USB through ports on hubs that incorporate status indicators to indicate the attachment or removal of a USB device. The host queries the hub to retrieve these indicators. In the case of an attachment, the host enables the port and addresses of the USB device through the device’s control pipe at the default address.

The host assigns a unique USB address to the device and then determines if the newly attached USB device is a *hub* or a *function*. The host then establishes its end of the control pipe for the USB device using the assigned USB address and endpoint number zero.

If the attached USB device is a hub and USB devices are attached to its ports, then the above procedure is followed for each of the attached USB devices. Alternatively, if the attached USB device is a function, then attachment notifications will be handled by appropriate host software.

When a USB device has been removed from one of a hub’s ports, the hub will disable the port and provide an indication of device removal to the host. The relevant USB system software must handle this indication. Note that if the removed USB device is a hub, the USB system software must handle the removal of the hub as well as any USB devices that were previously attached to the system through the hub.

‘Enumeration’ is the name given to the allocation of unique addresses to devices attached to a USB bus. Because USB allows devices to attach or detach from the USB at any time, bus enumeration is an on-going activity for the USB system software. Additionally, bus enumeration includes the detection and processing of removals.

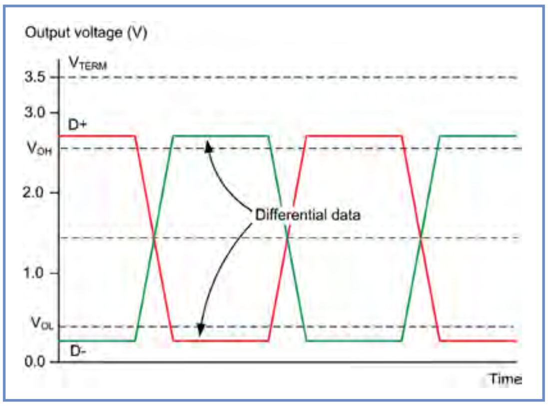


Fig.8.13. Signal levels present in a USB interface

Table 8.4 USB pin connections (see Fig.8.12) and conventional colours

Pin	Function	Colour
1	V_{BUS}	Red
2	D–	White
3	D+	Green
4	GND	Black

As mentioned earlier, USB employs two differential data lines (D+ and D–) and two power connections. CMOS buffers are normally used to drive the relatively low impedance of the USB cable and the signal voltage present on the D+ and D– must be kept within the ranges shown in Fig.8.13. Note that the terminating voltage (logic high) should be within the range 3.0 to 3.5V.

Detection of device connection is accomplished by means of pull-up and pull-down resistors placed respectively at the input/output of a port. USB pull-down resistors normally have a value of 15kΩ, while pull-up resistors have a value of 1.5kΩ. An interface adapter like that shown in Fig.8.14 can be extremely useful if you need to convert USB signal voltages to TTL-compatible signals (see Fig.8.14).

Loopback testing
The technique of loopback testing can be useful if you need to test a serial data interface; and is accomplished by looping back the transmitted data (TXD) back to the received data (RXD) line. Fig.8.15 shows the connections required to carry out a loopback test on an Arduino Uno microcontroller. When

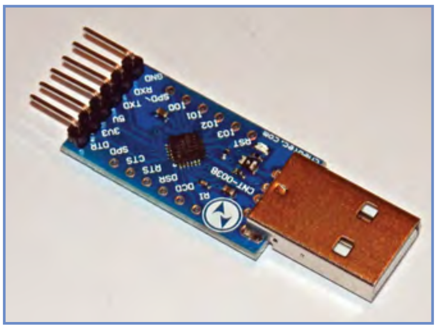


Fig.8.14. A low-cost TTL-to-USB interface

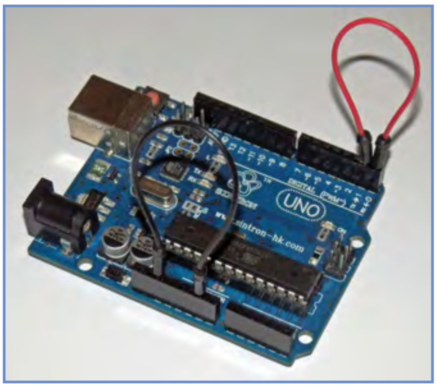


Fig.8.15. Carrying out a loopback test on an Arduino Uno microcontroller. The RESET line must be connected to GND (black link wire) and the TXD line to the RXD line (red link wire)

the links have been made the board is connected to a host computer via the USB interface and the serial monitor is then started, as shown in Fig.8.16. A string of ASCII text characters is first entered in the serial monitor before clicking on the Send button. The serial data then makes the round trip (by USB and RS-232) and is finally sent back to the host where it appears in the received data window. The data link can usefully be tested at different baud rates (in this case we have selected the fastest bit rate of 115200 baud).

Gearing up: Digital test equipment

Depending on the complexity of the circuit, digital test gear can be as basic as a hand-held logic probe and a few simple accessories. For more sophisticated logic, such as microcontrollers and microprocessors, a digital storage oscilloscope (DSO) is a useful acquisition (see Part 2). A DSO will allow you to capture a sample of data and then display it for detailed analysis at some later time. Data may be captured on a continuous basis or a trigger event selected in order to initiate data capture (note that it is possible to capture data both before and after a trigger event).

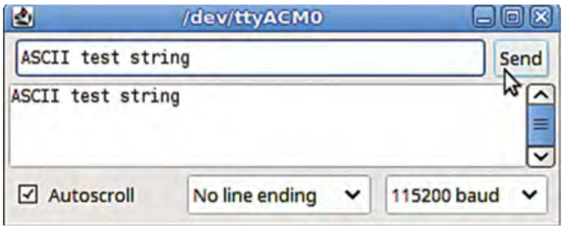


Fig.8.16. The serial monitor window showing the transmitted and received data

Fig.8.18. Captured digital data using LabNation's SmartScope software

Fig.8.17 shows how an external USB 'scope can be used to monitor the signal lines on a Node MCU microcontroller. The resulting display (captured and stored for analysis) is shown in Fig.8.18). If you need to debug more complex microprocessor-based systems on a regular basis then a dedicated logic analyser can be a useful investment. Unfortunately, such instruments can be rather expensive, but they do become available from time to time both second-hand and from on-line auction sources. Fig.8.19 shows a vintage SA3 Logic Analyser that can capture 40 data channels at a rate of 10 million samples per second. Instruments like this can often be purchased for as little as £50.

If you are working on a strictly limited budget it is still possible to enjoy logic analysis by using a low-cost USB bus interface like that shown in Fig.8.20. This handy gadget provides you with eight TTL-compatible input channels and is designed for use in conjunction with computer-based data-capture software,

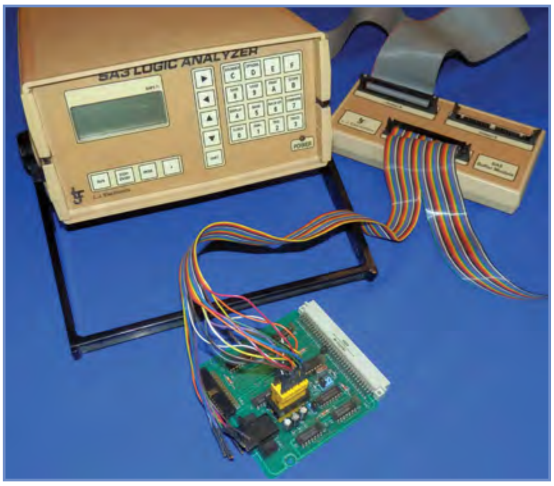


Fig.8.19. Vintage 40-channel SA3 Logic Analyser

Fig.8.17. A USB digital storage 'scope with auxiliary digital inputs connected to digital I/O lines on a Node MCU microcontroller

Get it right when when carrying out digital measurements

- When using a logic probe or pulser, take care to avoid short-circuits on adjacent pins or tracks
- When using a logic probe or pulser, ensure that you have connected the supply leads correctly and that the supply voltage is the same as that used on the logic that you are testing
- Before making measurements on logic circuits it is always worth checking that the supply voltage(s) are within the expected range (a low, high or missing supply voltage can produce misleading results)
- Always observe anti-static procedures when working on logic devices and particularly when removing and replacing them from a circuit board
- When using an oscilloscope to observe logic signals, set the input to DC and always use a compensated ×10 probe to minimise loading on the circuit under investigation (see *Teach-In 2018: Part 2* for details)
- If a bus line indicates an indeterminate state (ie, when neither logic 1 nor logic 0 is indicated when using a logic probe) it may be useful to momentarily pull the line high or force it low, and note the changes produced.

Fig.8.20. Ultra-low-cost 8-channel USB logic analyser

such as sigrok PulseView (from: **https://sigrok.org/wiki/PulseView**).

In addition to equipment for capturing and analysing logic signals, a variety of accessories will help you make effective connection to the circuit or system under examination. Fig.8.23 shows a typical selection, including adapters, a line monitor, probes and IC test clips. Finally, Fig.8.24 and 8.25 respectively show how a breakout boards and IC test clips are used in typical measurement situations.

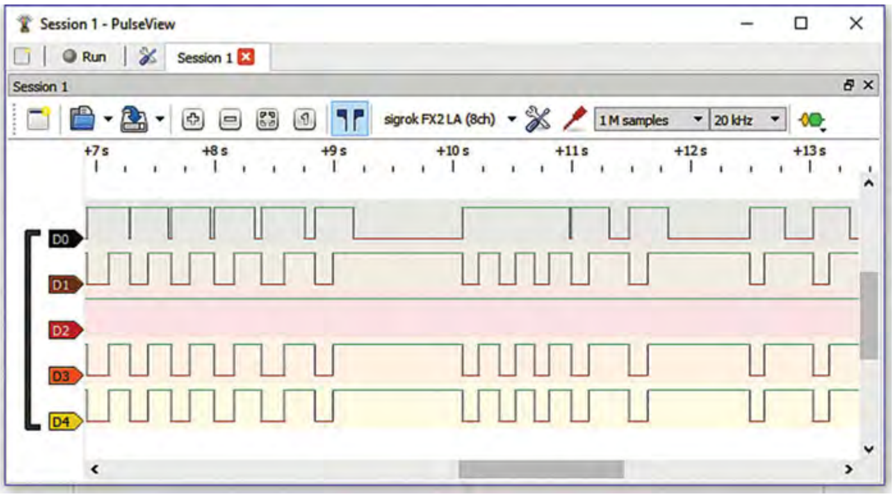


Fig.8.22. Using PulseView to capture and analyse data from a Node MCU microcontroller

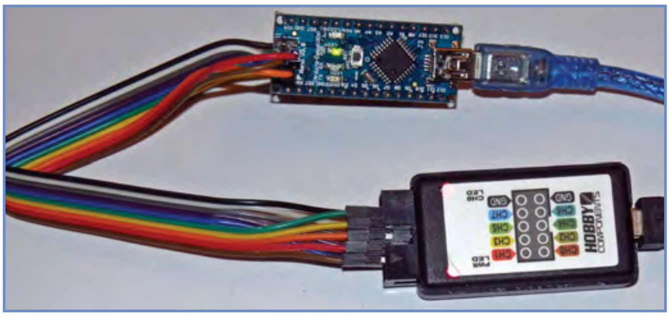


Fig.8.21. Connecting the low-cost logic analyser to a Node MCU microcontroller

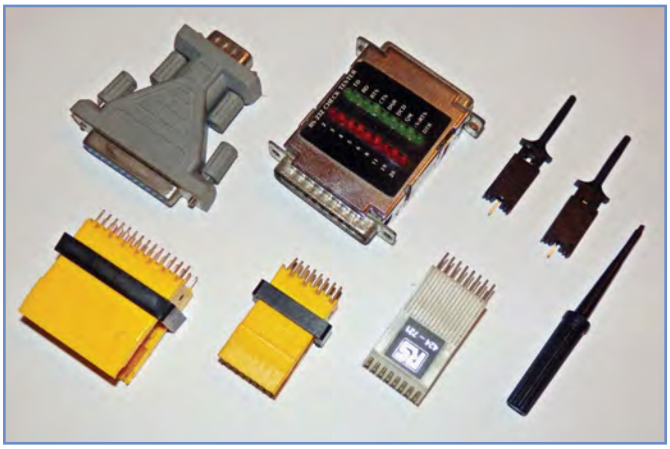


Fig.8.23. A variety of digital test gear accessories, including a 9-way to 25-way serial adapter, RS-232 line monitor, and various probes and IC test clips

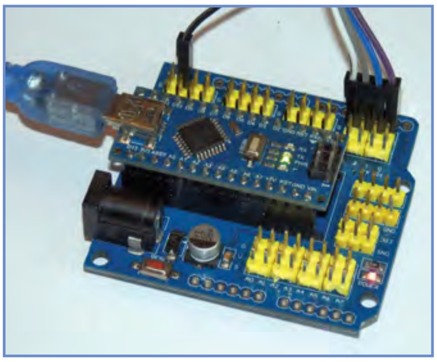


Fig.8.24. Using a breakout board to check logic signals present on an Arduino Nano

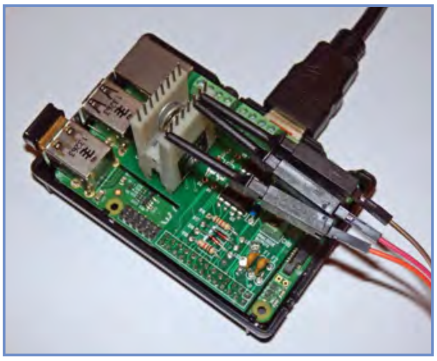


Fig.8.25. Using an IC test clip to check logic signals present on a Raspberry Pi expansion board

Test Gear Project: A simple logic probe

Our simple logic probe will provide you with a handy device for observing digital signals. Despite the lack of a pulse-stretching facility (see earlier) it is still possible to differentiate between static and clocked logic signals and make a rough assessment of duty cycle and mark-to-space ratio. This can be useful when it is necessary to determine whether a logic line has become stuck

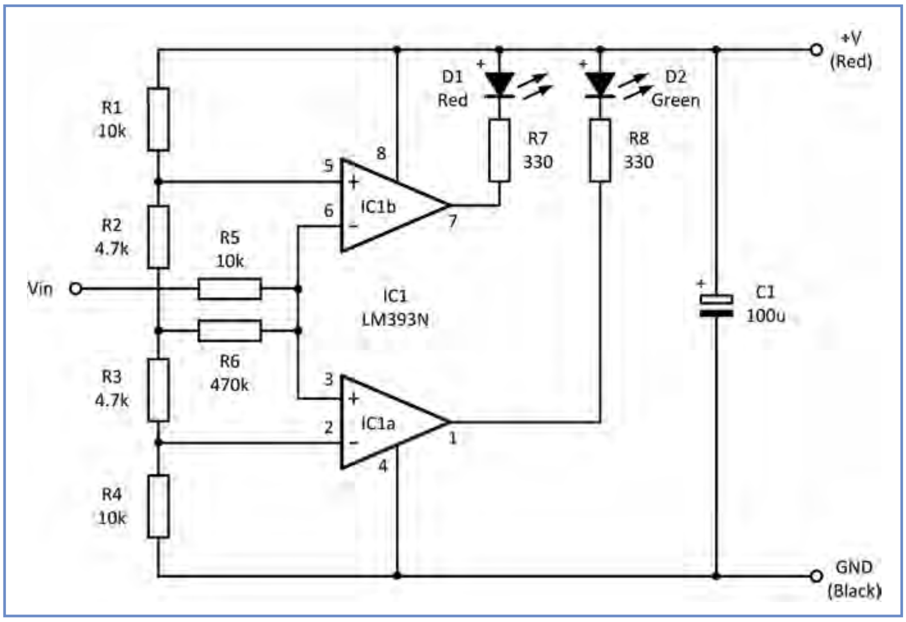


Fig.8.26. Complete circuit of the simple logic probe

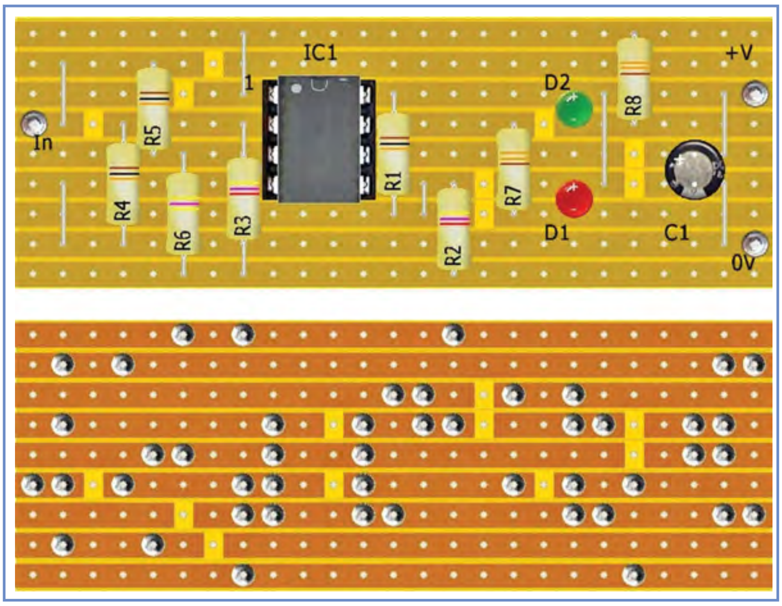


Fig.8.27. Stripboard layout of the simple logic probe

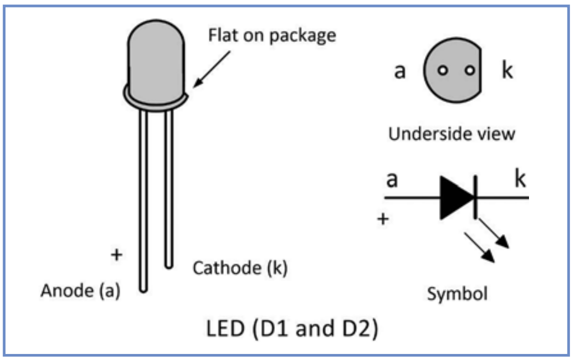


Fig.8.28. LED pin connections

Fig.8.29. Internal wiring of the simple logic probe

Fig.8.30. External appearance of the simple logic probe

or static (ie, permanently held at one or other logic level).

The complete circuit of our *Test Gear Project* is shown in Fig.8.26. The circuit comprises an LM393 dual comparator (IC1) and two LED indicators (D1 and D2) that indicate the state of the probe tip. If neither indicator is illuminated the probe is indicating a floating, indeterminate or tri-state condition.

You will need

Perforated copper stripboard (9 strips, each with 25 holes)
ABS logic probe case
Short length of twin insulated cable (see text)
2 insulated crocodile clips (black / red)
3 0.040-inch terminal pins
1 Miniature DPDT toggle switch (S1)
1 LM393N 8-pin DIL dual comparator
1 5mm red LED (D1)
1 5mm green LED (D2)
3 10kΩ resistors (R1, R4 and R5)
2 4.7kΩ resistors (R2 and R3)
1 470kΩ resistor (R6)
2 330Ω resistors (R7 and R8)
1 100µF 16V radial electrolytic (C1)

Assembly is straightforward and should follow the component layout shown in Fig.8.27. Note that the '+' symbol shown on D1 indicates the more positive (anode) terminal of the LED. The pin connections for the LED are shown in Fig.8.28. The reverse side of the board (NOT an X-ray view) is also shown in Fig.8.27. Note that there's a total of ten track breaks to be made. These can be made either with a purpose-designed spot-face cutter or using a small drill bit of appropriate size. There are also seven links that can be made with tinned copper wire of a suitable diameter or gauge (eg, 0.6mm/24SWG).

When soldering has been completed it is very important to carry out a careful visual check of the board as well as an examination of the track side of the board looking, for solder splashes and unwanted links between tracks. The internal and rear-panel wiring of the test signal source is shown in Fig.8.29. Finally, the PCB should be placed in the logic probe case (it should fit snugly inside the case with two holes drilled in the removable panel for D1 and D2). The probe enclosure used for the prototype was a Teko LP1 Probe Case, measuring 145×30×21mm and available from Rapid Electronics (Order code 31-0335).

The probe tip should be connected to the input terminal pin via a short length of insulated wire. The supply is connected using a twin insulated lead (400mm is ideal) terminated with red and black crocodile clips and soldered to the +V and 0V pins on the circuit board (see Fig.8.29).

Testing

Before use, it is important to test the logic probe, ensuring that the logic levels are correctly identified. Connect the supply leads to a 5V DC power source as shown in Fig.31. The probe tip is taken to the slider of a 1kΩ potentiometer that can then be adjusted to produce an input voltage of between 0V and +5V. The voltage at the probe tip is indicated using a digital multimeter and the voltage range for logic 0 (green LED illuminated) and logic 1 (red LED illuminated) can then be observed.

If the logic probe is working correctly the ranges should be 0V to 1.7V for logic 0 (green) and 3.3V to 5V for logic 1 (red). Note that neither LED should be illuminated for input voltages between about 1.7V and 3.3V. If this is not the case, check the orientation of IC1, the

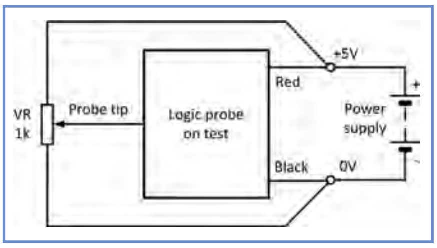


Fig.8.31. Test circuit for the simple logic probe

Table 8.5 Threshold voltage levels for low and high logic states

Supply voltage	3V	3.3V	5V	9V	12V	15V
Low threshold	1.0V	1.1V	1.7V	3.1V	4.1V	5.1V
High threshold	2.0V	2.2V	3.3V	5.9V	7.9V	9.9V

polarity of D1 and D2, and the circuit board wiring (checking that the track breaks and links have all been made correctly). Table 8.5 shows the typical threshold voltage levels when the logic probe is used with different supply voltages. Note that, under no circumstances should the supply voltage be allowed to exceed 15V.

Next month

In next month's *Teach-In 2018* we will bring the series to a close with some advice on designing and building your own test gear. We will also include an index to all previous parts of this series.

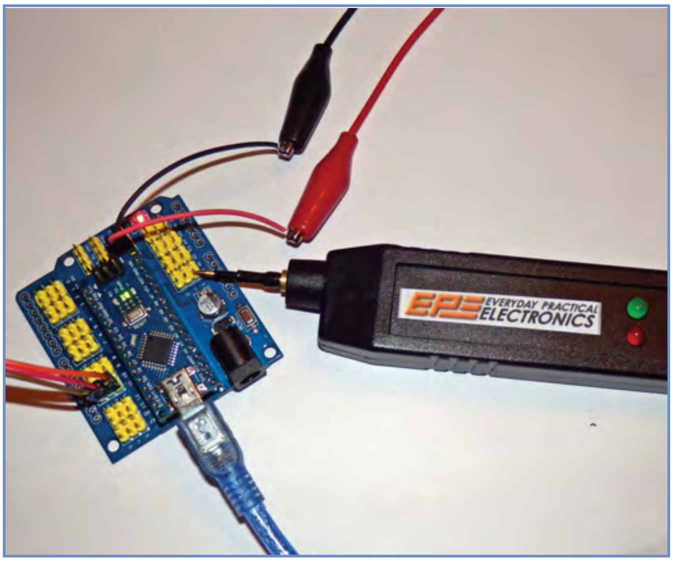


Fig.8.32. Typical supply lead connections. On this Nano breakout board the red crocodile clip is taken to '3V3' and the black is taken to 'GND'

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NET WORK

by Alan Winstanley

Alexa – make me some money!

BACK IN 2013, an electronics inventor called Jamie Siminoff appeared on the US TV show 'Shark Tank', a program similar to the BBC series *Dragon's Den*, in which embryonic business owners showcase their brainchild to a group of angel investors in the hope of winning some seed capital. He demonstrated an early version of an Internet-enabled doorbell called 'DoorBot', which had a built-in video camera that allowed owners to see, and speak with, visitors on their smartphone. In theory, DoorBot would help ensure homeowners never missed a visitor or delivery, especially when occupiers were away from home. Siminoff sought $700,000 for 10% of his business, but most of the 'sharks' rejected the concept. Siminoff turned down the only offer on the table and he walked away empty handed.

That was five years ago, but after several rounds of funding the product has undergone radical quality improvements to become the Ring video doorbell, as advertised on TV during Christmas 2017. None other than Amazon snapped up the entire product range a few weeks ago, reportedly for an eye-watering $1.1bn. Amazon seeks to add Ring video doorbells to its range of Internet-capable home control devices, and Ring is now part of the Alexa Fund portfolio, Amazon's own venture capital fund that supports projects that could benefit from voice control technology – specifically, the Alexa Voice Service. Since Amazon has now launched Echo Show, a £200 wedge-shaped smart speaker with built-in LCD screen and camera, it made perfect sense to integrate a video doorbell into Alexa's ecosystem by offering an Alexa Skill. You can now 'ask Alexa' to 'show you the front door'. Ring also produces an interesting range of security cameras, including rechargeable and solar-powered ones, promising a marriage with Alexa that is made in heaven.


Apple's £319 Homepod missed a Christmas launch and so came to the smart-home party rather late – its sound quality has been praised, less so its Siri-based 'smarts'

Amazon's £120 Echo Spot boasts a circular LCD screen and a control interface that is Alexa-powered

Amazon might perhaps go even further, scanning images of parcels or delivery staff into a Ring camera to take smart shipment tracking to the next level. Just as Google swallowed up Nest and its smart thermostats, doorbells, security cameras and more, Amazon's latest acquisition points the way to a diversifying range of Alexa-aware home security and video doorbell gadgets that will enhance life in the networked home and become second nature to today's generation.

If smart doorbells and smartphones just aren't for you, readers might consider a wireless doorbell system instead, such as Honeywell's modern long-range wireless door chimes, push buttons and sender units that are readily available online. More details at: **https://livewell.honeywell.com/en/doorbells/**, or consider the Lloytron MiP range: **www.lloytronuk.co.uk/mip/**

Maplin's sinking sands

During a nasty snowstorm last month, I called into a small, traditional menswear shop in search of some winter clothing. The shop resembled something from 'Hoarders', with every square inch of floor and wall space laden with stock, leaving me with nowhere to stand. The owner popped up from nowhere and we got down to business. I could try on a range of gloves in colours, size and (most importantly!) fur linings until I found what I was looking for. It was pricey, but the owner proudly exclaimed that his retail prices were still lower than the official online ones. Despite the crazy stock control, this half-century-old business was still trading reasonably well in the Internet age, offering personal service on goods that you could instantly take away with you; it only stayed in business though because customers appreciated the service and went out of their way to support him.

So it was with great sadness that I learned of the looming demise of Maplin, the UK's largest high-street retail chain of electronic technology and gadgets, and a brand familiar to legions of *EPE* readers. At the time of writing, the stores continue to trade normally after the business collapsed into administration at the end of February. Maplin has passed through several owners' hands in its chequered history and

the loss-making retailer has struggled in the face of faltering demand for ephemeral technology lines that churn over far too quickly, choosy consumers and adverse exchange rates (a problem for any trade sector reliant on Chinese imports, which are priced in US$). Last Christmas, I saw how local Maplin stores were busy pushing Google Home and Amazon Echo smart speakers (the Apple Homepod had been delayed and missed the rush), the Ring video doorbell and all manner of home security and other gadgets. However, I could tell that consumers were unsure and business was far from brisk: the nearest store was sparsely populated despite Maplin's valiant attempts to draw in the crowds. Super-rapidly changing technology products and flaky consumers seem to be a toxic mix. It's a sad but inescapable fact that the Internet has skewed buyers' *modi operandi*, and today's retail market has been distorted out of all recognition by online shopping bringing millions of products and services to buyers' doors almost effortlessly.

Maplin stemmed from a highly efficient mail order business that sold kits and a few parts for projects in popular hobby electronics magazines in the 1970s. This gave the firm a good head start in 'arm's length' retailing and it has worked hard to embrace the web, search engine marketing, social media and all the rest of it. Hindsight is always 20/20, but the firm probably expanded too optimistically into bricks and mortar stores, over 200 of them, any number of which are now under threat.

A changing hobby

I can hear readers saying that Maplin isn't what it used to be, which is true: but flogging penny resistors or odd bits of stripboard won't pay the wages or the shop rent. Electronics as a modern hobby has changed since Maplin's glory days preceding the 1990s, and cheap and powerful microcontrollers now form the core of projects. Most discrete parts, tools and equipment can be bought at dirt-cheap prices online. Indeed, our current series *Using Cheap Asian Modules* shows how easily a cheap-and-useful, ready-built circuit module can be bought and imported this way.

Could Maplin catch the tidal wave of a new era of technology? Diversifying into higher-margin smart speakers, IoT gizmos, home CCTV and electricals as well as Raspberry Pi or Arduino made sense on paper, but again, almost everything could be bought online instead and judging by the lack of visitors to local stores, the demand for goods did not seem to materialise. The consumer's habit of 'showrooming', or testing out a product's look and feel in-store before going home and buying it more cheaply online, hasn't helped retailers in general either. As more stores are gradually driven out of business by consumers' changing habits, buyers will find they have less choice than they enjoyed before. Having grown up with Maplin since it started trading from a spare bedroom in the 1970s, we at *EPE* hope that Maplin's administrators find a workable solution that will keep the brand alive for future hobbyists to enjoy.

More smart talk

Two more smart devices have arrived on the UK market, including the pricey and overdue £319 Apple Homepod smart speaker working with Siri, which seems to have had something of a soft launch. Early reviews are impressed by its high audio quality, but Homepod is probably better at playing Apple Music or tracks purchased from the iTunes store (in the same way an Echo can play tracks or CDs purchased from Amazon), rather than using it as a general voice-activated domestic helper like Google Home or Echo. Amazon UK has also released the Echo Spot, a £120 Alexa-powered product with LCD screen and camera in a round form factor about the size of a large apple (see *Net Work*, December 2017). This, coupled with Amazon's new doorbells and security systems shows how the battle to take control of the smart home is becoming more hard-fought than ever.

Practical DSP – Part 2

HELLO again! In this second article on DSP (digital signal processing) with Microchip's dsPIC33 microcontroller, we move to installing the development environment and support files to enable us to create our application. The development environment consists of the MPLAB X IDE (integrated development environment), the XC16 C compiler (the dsPIC33 has a different code compiler to the usual PIC16 and PIC18 processors) and an example application from Microchip to help us get started.

Free tools

All these software tools are provided free by Microchip. The XC16 compiler is available with a 'free or paid-for' model. It's the same download, but you can select which version you are using during installation by either selecting 'operate in free mode' or entering a license code. To Microchip's credit, the free version is plenty powerful enough, and we will not miss the additional features – mainly code space optimisation – in our projects. The optimisation aspects come in handy when you want to reduce code size to fit the code in a device with less Flash memory. That's important when you are selling tens of thousands of products; but for us, we can simply pay a few pounds more and target a larger memory chip, as we have done. As an additional bonus, the DSP library functions – the code that we care about being optimised – are supplied pre-compiled and therefore pre-optimised.

For this article series we are using the latest version of Microchip's development tools, namely MPLAB-X v4.10. If you would like to follow along then we suggest you download this too. The code shown here is compatible with MPLAB-X, not the old MPLAB IDE. If you are downloading the IDE, then allow installation with the offered defaults and allow installation of additional device software – this step adds drivers for the PICkit 3 and related debuggers.

Once installation completes, your web browser will launch and open on several pages. Click on the **MPLAB-XC compilers tab**, scroll to the bottom and select **Downloads**. Now click on **MPLAB XC16 Compiler v1.33** to download the installer. Once the download completes, install that too, accepting all the default options. Once the installation completes you can go ahead and run the MPLAB X IDE by clicking on the desktop icon. Do note that it's the IDE we run, not the similarly named 'IPE', which is the programming tool.

Exploring the DSP library functions

The **Help** menu for the XC16 Tool chain hints at the existence of the DSP library, and tells you to go search for the documentation in the installation directory. We eventually found the DSP library and documentation, but it appears dated – the user guides refer to it as the **C30 DSP library**, which is a different device family to ours. Buried within the source code zip file however, an updated documentation states it is the **XC16 DSP Library**. So, fingers crossed we will not come unstuck with our choice of processor.

As is normal with documentation about library functions, the guides tell you what the functions do, not how to use them. Thankfully, example code is provided to show the use of the library in some realistic contexts, and these example applications are always a vital source of knowledge. Here again we come have a problem – the FFT example project **CE018** is very old, not modified since 2007 and designed to work with the old and incompatible MPLAB v7 IDE, and the C30 compiler (now retired and replaced by XC16). So, potentially, we now face a significant porting job to make it build under MPLAB-X.

Thankfully, a search on the Internet revealed that a more recent example project has been created by Microchip, named **CE482**. This was last updated in 2015, targets MPLAB-X and the XC16 compiler, and specifically refers to the dsPIC33 processor family. With luck, our workload has been dramatically reduced.

In situations such as this, where you are starting on a new project, it's advisable – highly advisable – to take small steps, validating each step as you go. With this approach in mind, we started by downloading the **CE482** example project from Microchip, opened it within MPLAB-X IDE and attempted to build it with no modifications. It was with some great relief that this build completed successfully 'out of the box'. A quick scan of the build process output messages shows that the total code space used is 3714 bytes, and 3672 bytes of RAM. Very small, which means our choice of processor should be able to hold the DSP library functions required with plenty of room for future application code – looking good!

The next step is to change the project to target our selected processor and try building again. We are not worried about running the code on our board yet; in fact, we have not even looked to see exactly what the code does, let alone whether it will run on our minimalist hardware. Remember, it's baby steps, one change at a time. We creep slowly towards having code run on our board. The example application has been designed to run on Microchip's Explorer 16 board, an extensive but pricey solution, more than ten times the price of our hardware. With luck (and a little forward planning!) the code will be easy to 'port' to our board.

Porting

The process of taking a program written for one microcontroller and moving it to another is called 'porting'. Just how complex this task becomes is down to several factors:

- How dependent the application is on a microcontroller's specific features
- The level of differences between the microcontrollers
- How well the application was written
- The clarity of the porting documentation

The first point above is covered; we are dependent on the DSP instructions in the processor, specifically within the Microchip-supplied DSP library that our

application links to during compilation. As that library supports our processor, and the example application does not connect to an external signal source, we should be good.

The second point is reasonably well covered because we are sticking with a processor in the same family, with a large amount of code and data space, so we don't expect any (nasty) surprises.

The third point is hard to judge, so we proceed by checking the porting guidelines and dive right in. The **Readme** file for the example application hints that the process will be simple:

Change device selection within MPLAB IDE to other device of your choice by using the following menu option:

MPLAB X >> Configuration drop-down option >> <Listed Device Configuration>

And that's it. Great – this is going to be simple!

Uh oh, not so fast – a look at the options provided in the drop down allows for only three processor options, and they do not include our processor (see Fig.1). What do we do now? A quick look at the files included in the project shows that there are several files specific to each processor offered in the drop-down list. As should be expected, each processor has a different amount of memory and different configuration bit settings that need to be accounted for when building any application. So, there are different files for each one. Fortunately, one of the processors listed – the dsPIC33EP256GP506 – is very similar to ours. It just has more pins and less memory.

We now have two options: 'hack' the settings for the dsPIC33EP256GP506 configuration files to match ours, or create a new configuration specifically for ours and do this properly. After much soul searching we chose the latter, as this way we are extending the Microchip example rather than creating a hard-to-maintain modification.

The porting process starts by duplicating the configuration settings, an option provided within the IDE. We rename the configuration to **dsPIC33EP512GP502**, change the debugger to **PICKIT3**, and select the correct processor type. Next, we copied the source code directory from **src\system_config\exp16\dspic33ep256gp506** to a directory called **src\system_config\epe\dspic33ep512gp502**.

Finally, we added the newly copied-over source file to the list of source files within the project itself. Performing a build of the project resulted in success, so that's the porting over. It took just three hours of puzzling, which is quick, even for a simple project such as this.

The next step is to fully examine the source code of the Microchip example before we build our own, to understand the context of how to call the DSP library functions. If you would like to follow these articles along fully, go ahead and download the complete set of project files, which can be found on the current issue's monthly web page at **www.epemag.com**. (In case you are wondering, the original license from Microchip permits modification and redistribution.) We will be expanding on these files over the coming months, so make sure you check out each month's downloadable files from the website.

Inside the example application

Let's go up a level and review where we are. The point of the exercise this month is to obtain a configuration of the DSP library that will run with our processor, and to understand the context of how the DSP library functions should be called. The help documentation within the IDE explains what the functions do, but not how to string them together. With luck, the example application supplied by Microchip should do this, so let's look.

The bulk of the code is held in the file **main_fft_example.c**, located in the directory **src\system_config\epe\dspic33ep512gp502**. Open the file in a text editor if you would like to follow along.

As a side note, IDEs such as MPLAB-X are not the best text editors when all you want to do is quickly view the contents of a text file. Notepad, supplied with Microsoft Windows, is acceptable, but very basic. Our recommended text file editor is Notepad++, a free and fully functional text editor. It opens quickly and does not fill the screen with clutter. You can download it from the creator's website here: **notepad-plus-plus.org**

The first section inside **main_fft_example.c** is the list of configuration bits settings. These are important because they define, among other things, the settings your microcontroller uses for the main system clock – and since this a different processor, it is almost certainly wrong. Correcting this will just require comparing the configuration settings section of both datasheets. We will come back to this section later.

Next, in the file, some variables are declared. Thankfully, just six variables, which will help simplify our understanding. Let's look at the four important ones.

`fractcomplex sigCmpx[]`

This is an array that will hold the input data to the FFT function call. Note the unusual data type, `fractcomplex`. This is a special data type created by Microchip for use with their DSP library. A variable of type `fractcomplex` consists of two parts, the real part and the imaginary part. These are 'complex numbers', not simple integer or floating-point values.

Complex numbers are used in the FFT calculation because the signal is composed of many different frequencies varying in both amplitude *and* phase – complex numbers enable us to represent both quantities. A fuller explanation of complex number theory is beyond the scope of this article series, but it's fine for us to simply ignore this fact, and just take care to follow the data types being used, and how they are translated back to useful and meaningful values.

When the FFT function is called, the resulting frequency data is returned in `sigCmpx`, so a second output array is not required.

`fractcomplex twiddleFactors[]`

This unusually named variable is actually a now-standard term used within DSP algorithms. This is an array of constants that are used within

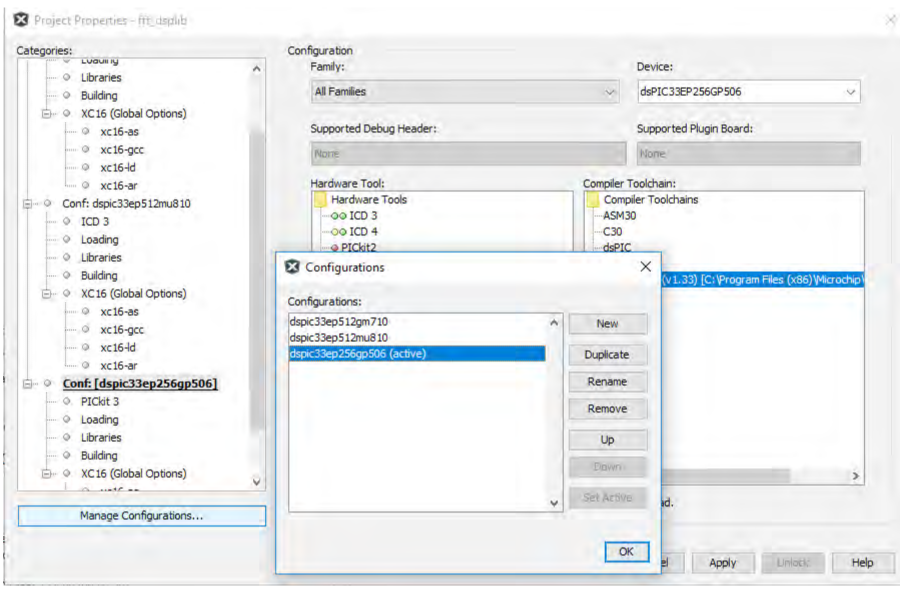


Fig.1. Changing configuration in the IDE

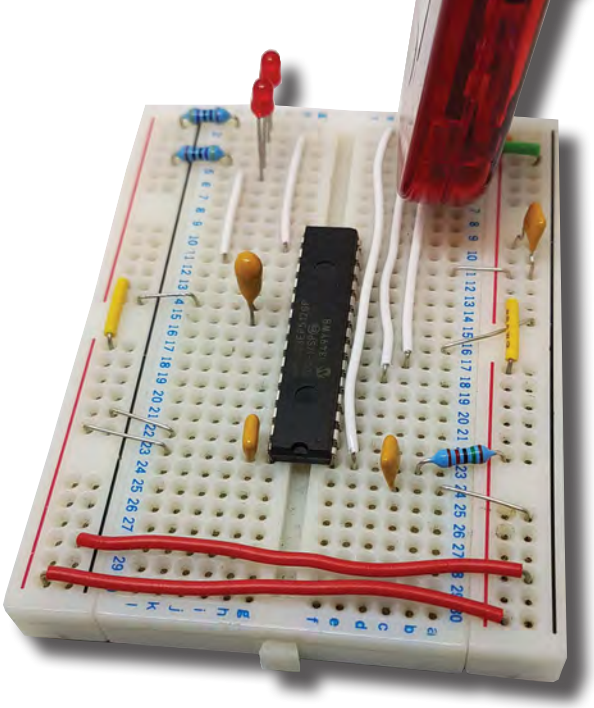


Fig.2. Completed breadboard

the FFT function itself. The values change for FFT algorithms of different input sizes, but for a fixed input data size, the constants are the same – so it is convenient to pre-compute these before use. The Microchip example uses an FFT size of 512 samples, and we will stick with this, to save having to understand how to calculate new twiddle factors. We will talk about the implication of using a 512-sample buffer later, as the length of the buffer used impacts how long it takes to perform the FFT calculation, and the resolution of the frequencies obtained.

`fractional input[]`

This is an array in which the pre-computed example input signal is stored. In our example code the contents, held in the file **inputsignal_square1khz.c**, define a 1kHz square wave signal. Notice that again, this datatype is a Microchip defined one. It is used to represent a special form of non-integer values that are encoded into a 16-bit integer variable. In next months article we replace the contents of this array with samples taken from the ADC peripheral. The use of a fractional datatype is very helpful because the ADC peripheral supports outputting data in a fractional format – so once the ADC is correctly configured, we will not need to modify the data coming from the ADC before placing in this buffer. That will save lots of processing time.

`fractional output[]`

This array is used to store the output of the conversion from the FFT function's complex data output and a squared magnitude representation, performed by the function `SquareMagnitudeCplx()`. The output buffer holds the data that we will use in our subsequent calculations. We make use of the square of the magnitude rather than calculate the magnitude itself, as the square-root function required to do this is very time consuming. For our applications, where we are only interested in finding the peak frequencies, it's not necessary to perform the square root computation. We used exactly the same technique in our speed camera detector project back in January 2005 to provide a fast display update.

That short list of variables is followed by a simple `main()` function, which performs all the function calls. It's actually quite straightforward, and the list of functions called is very short:

- Copy the input data to the `sigCmpx`
- Call the FFT
- Compute square magnitude
- Find the peak frequency

A point of interest is how the fractional input array is copied to the `fractcomplex sigCmpx` array. Two steps are involved; the imaginary component in `sigCmpx` is set to zero, and the real component is set to half of the input data value. This is a requirement of Microchip's FFT function and is mentioned in the DSP library help file. Although a fractional variable can store values between –1.0 and +1.0, the FFT function requires that data be scaled down to between –0.5 and +0.5. A simple shift-right operation performs this.

Running the sample application

Returning to the configuration bits, we selected the internal RC oscillator as the source, and enabled the PLL (phase-locked loop) to multiply the RC oscillator's output to 140MHz, giving us the highest processing speed of 70 million instructions per second. We also added code within the `main()` function to toggle our LEDs, enabling the use of an oscilloscope to measure the execution time of the FFT conversion. The breadboard setup, now connected to the IDE and ready to go, is shown in Fig.2, and the schematic in Fig.3. Note that we do not have an external power supply – the PICkit 3 can provide the power itself.

We were delighted to find that the full process of copying data, FFT, square magnitude and peak detection took just under one millisecond. This is fantastic, because even when we add the time required to perform 512 ADC samples, we will be running fast enough to be able to update a display several times a second. So, our tiny microcontroller is quite a powerful beast!

Final thoughts on FFT

Although we have discussed the magic of the FFT (Fast Fourier Transform), we've not yet spoken about the relationship between signal sample rate and FFT block size, and how these two relate to the values that are output by the FFT. Hopefully, with the aid of Fig.4 we can explain that.

At the top of Fig.4 we have our input signal – in this case, a nice simple sinewave. This is being sampled at a periodic interval. Eight samples are presented to the FFT algorithm, so the block length in this example is eight. (In our code we are using 512 samples, but this example shows just eight for clarity.) With eight samples, our sampling window (T_w) is eight times the ADC sample rate.

The FFT provides eight outputs – the same number as the number of inputs. Each output is a complex number,

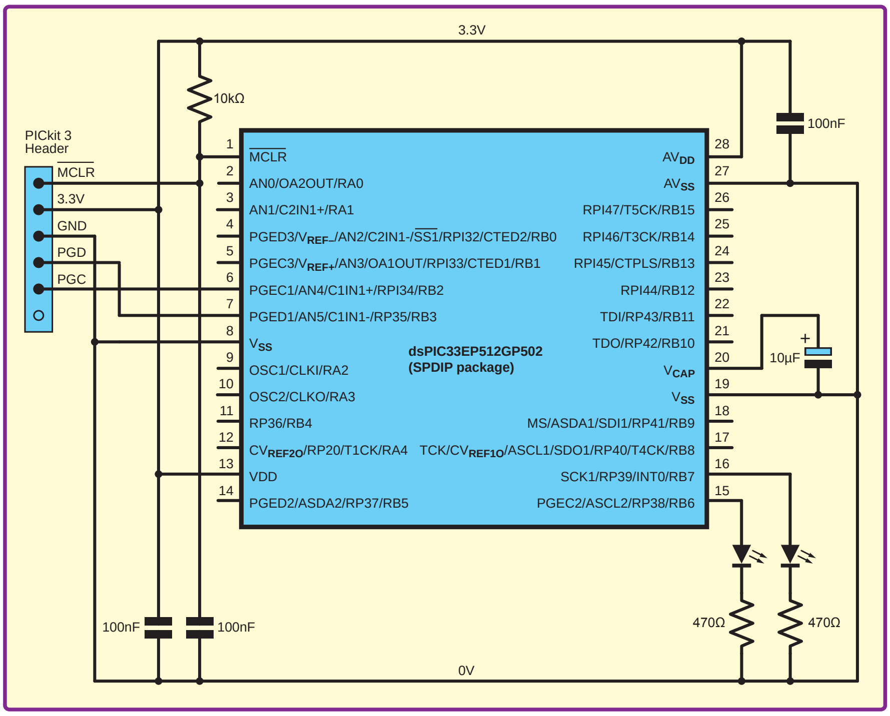


Fig.3. Breadboard schematic

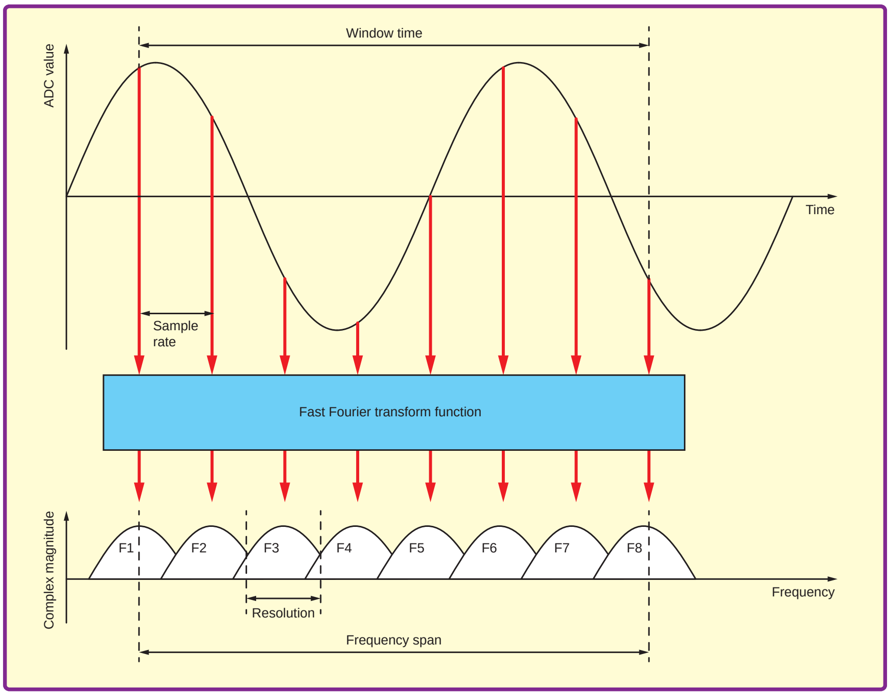


Fig.4. FFT conversion

denoting a specific frequency, and its magnitude. Each of these outputs is called a 'bin', as it contains the magnitude of a range of frequencies, the range being the resolution of that bin. Each bin spans a frequency defined by $1/T_w$.

What about trying a real-world example? Well, next month we will be listening to the sound of guitar strings. Let's make a rough assumption that the maximum frequency we are interested in resolving is 1kHz. If that is the maximum, then Nyquist's theory says we need to sample at twice that rate, so our sample rate will be 2kHz. We have an FFT block size of 512, so our window size is 256ms. Given that the bandwidth of each bin is defined as $1/T_w$, each bin spans a frequency range of 3.9Hz. Probably good enough for tuning a guitar – but we are not experts, so do feel free to comment!

If better resolution is required, then clearly that means spending more time taking samples, and having a bigger FFT, which will take longer to compute. So, there is a delicate balance between accuracy and response time.

Next month

In next month's article we expand the circuit to include audio input from an electret microphone, so we can collect data from the real world, and attempt to create a guitar tuner. The choice of Electret microphone is largely uncritical; we are using the cheapest one available from Farnell – part number 2066501. Costing less than a pound and with free next day delivery, it is simply not worth rummaging around to find an old discarded microphone with unpredictable performance. We are not looking for high audio quality here, so the cheapest option will be perfectly acceptable.

The other circuit change we will have to add next month will be correctly scaling the input signal from the microphone to match the input range of the microcontroller's ADC. If the output from the microphone is a few hundred millivolts, then we will need to amplify it by a factor of 10 or more. To achieve this, we plan to make use of a novel peripheral integrated into the microcontroller – an op amp, shown in Fig.5. It would seem that Microchip have included this for exactly the purpose we are intending, to reduce the circuit component count by two components, a dedicated op amp IC plus resistor. Once again, we would not be using this if we were looking for a high quality op amp, but given our project needs, it should be spot on. This is the first time we have used an op amp integrated into a microcontroller, so next month is going to be exciting!

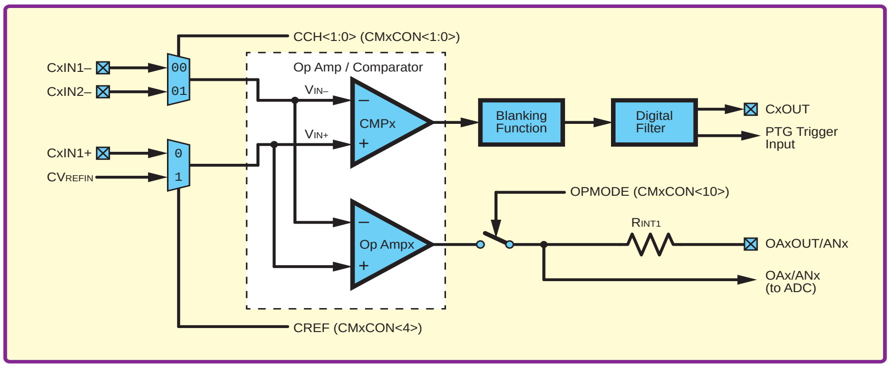


Fig.5. DSPic OP-AMP peripheral

If you are wondering how you will follow this article series along if you do not own a guitar – which most of us do not – fear not. We will be using a freely available audio signal generator program for the PC to create the guitar string sounds, available from here: www.ringbell.co.uk/software/audio.htm. And there are many different audio generators available, even for smart phones. So testing the circuit will not be difficult. We have several guitars available, so this project will be getting a genuine real-world test next month.

Source impedance and regulator stability

IN THE February and March *Circuit Surgery* columns I discussed some issues related to power supply design posted by **Tuurbo46/Rocket Ron** on the old *EPE Chat Zone* (and continued on *EPE's* new forum home on EEWeb: **www.eeweb.com/forum/**). Martin Walker responded to these articles with another forum thread on EEWeb, tagged with 'EPE Magazine' to highlight its relevance to *EPE*. He wrote: 'I have been enjoying the recent *Circuit Surgery* columns on power supplies and potential dividers in *EPE Magazine*. I love the way that Ian Bell tackles each subject, explaining the theory clearly from first principles using worked examples to aid understanding. I wondered whether *Circuit Surgery* would take a look at a subject that has always been a little fuzzy for me – switching power supply source impedance and how to mitigate for it. We all know the rule of thumb that the source impedance of the input to a voltage regulator must be lower than the negative input impedance of the converter (negative because the line voltage drops when more current is taken by the regulator) by a factor of at least 10 times. We also know that if the source has a high impedance (e.g., long wires from the source to the regulator) then this can be mitigated with some input capacitance, usually a large-value electrolytic. What's not clear to me is how big that capacitor should be for a given source impedance. Explanations range from the highly mathematical to the generic 'use a 470µF for everything'. Is there a way to work out a suitable minimum capacitance value, based on the estimated or calculated impedance of the source?'

First, thanks to Martin for his positive comments about the articles. Martin obviously has some familiarity with this topic, but we will step back a bit to explain some of the background, including basic switching regulator operation, before discussing some of the issues related to input capacitor selection.

The basic configuration, without any capacitors or filters added, is shown in Fig.1. A DC source provides a voltage (V_I) to the input of the regulator, which in turn delivers a regulated output voltage (V_O) to the load, R_L. V_O will remain (more or less) fixed for a certain range of source voltages – the input operating range of the regulator. V_O will also remain fixed as the load varies (again within limits), implying a variation in load current. In the context of Martin's question the regulator is assumed to be a switch-mode DC-DC converter.

As discussed in the February article, real voltage sources (e.g., batteries and mains-rectified and/or stepped down by a transformer) are not perfect, but can be represented by an ideal voltage source in series with a source resistance. Where a circuit is simplified to this form, it is known as a 'Thévenin equivalent circuit'. For sources such as batteries, R_S is referred to as the 'internal resistance'. Fig.1 shows the source/internal resistance for the voltage source and regulator (R_S and R_P respectively).

The regulator takes electrical power from the source, at voltage V_I and delivers it to the load at voltage V_O. The input power taken by the regulator is $P_I = V_I I_I$ (I_I is the regulator input current, see Fig.1). The output power delivered to the load is $P_O = V_O I_O$ (I_O is the regulator input current/load current). A real power supply will not deliver all of its input

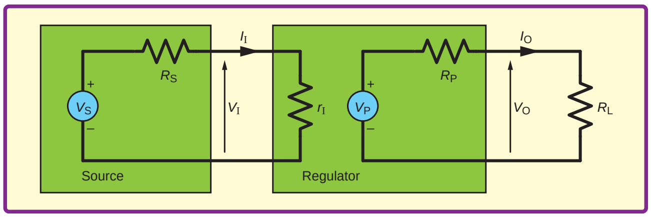


Fig.1. Power supply unit connected to a source and load

power to the load because it will require power to operate and some losses will occur. The ratio of input power to usable output power is the efficiency (η) of the regulator, where: $\eta = P_O/P_I$. Switching regulators can achieve 80% efficiency or better, but in the following discussion we will assume 100% efficiency to simplify things.

Fig.1 shows that the regulator has an input resistance; however, this is not a simple constant resistance. For an ordinary resistor, an increase in voltage across it will result in an increase in current through it, but *here*, because the regulator's power is constant (with constant load), increasing the input voltage *decreases* the input current. This opposite direction of change means that the effective input resistance of the regulator is negative (as mentioned by Martin). Furthermore, the input resistance of the regulator is not constant, so we refer to it as incremental or dynamic resistance and use a lower-case symbol, r_I.

Dynamic and negative resistance

For an ideal resistor, with resistance R, we have the well-known Ohm's law relationships $I = V/R$, $V = IR$ and $R = V/I$. If we plot a graph of current (I) against applied voltage (V) we get a straight line. The generic equation for a straight line is $y = mx + c$, where m is slope of the line and c is the value of y when $x = 0$. For a resistor, y corresponds to I and x to V. Zero voltage results in zero current, so $c = 0$ in the generic equation, leaving $y = mx$ or $I = (1/R)V$ – thus the slope of the line is $1/R$. A steep slope (large increase in current for a small increase in voltage) indicates a small resistance and vice versa.

The resistance of an ideal resistor does not change with voltage or current (the slope of the line of the *I-V* graph is constant). However, the resistance of other components is not always the same at different applied voltages. In such cases we cannot state a single value of resistance, but we can measure the slope of the line at any point on the curve of interest to find the (dynamic) resistance (r) at that point. We can do this by choosing two points close together on the curve (V_1, I_1) and (V_2, I_2) and then determine the slope at that point – the ratio of the change in the y value (I in this case) to the change in the x value (V in this case):

$$slope = \frac{1}{r} = \frac{\Delta I}{\Delta V} = \frac{I_2 - I_1}{V_2 - V_1}$$

Finding $1/slope$ gives the incremental resistance. The slope is most accurately determined if the difference between the two points is a small as possible. In the limit, the difference tends to zero and we write dV instead of ΔV. This takes

us to the mathematics of calculus and the definition of dynamic resistance as the differential of the voltage-current relationship:

$$r_d = \frac{dV}{dI}$$

If current decreases as voltage increases then the slope is in the opposite direction to that of a resistor and the resistance is *negative*. (Note that we can still use small differences or differential calculus to find the resistance at a point of interest on the curve.)

Regulator dynamic input resistance

A regulator contains a negative feedback control loop, which attempts to keep its output voltage constant; this will result in changes in input current if the input voltage changes. However, initially consider the situation where the load and input voltage do not vary. We have $P_O = V_O I_O$ and $I_O = V_O/R_L$, in which V_O and R_L are constant due to the unchanging situation, so I_O is also constant and hence P_O is constant. Assuming 100% efficiency, the input power must also be constant and equal to P_I. Therefore $P_I = P_O = V_I I_I = V_O I_O$. From this we can also write $I_I = P_I/V_I = P_O/V_I = V_O I_O/V_I$. With everything constant we can consider the input resistance as a standard voltage/current ratio – this is referred to as the DC or static input resistance of the regulator, $R_{I,DC}$

$$R_{I,DC} = \frac{V_I}{I_I} = \frac{V_I^2}{V_O I_O}$$

Here, I_I was substituted with $V_O I_O/V_I$, as derived above. As an example, consider a DC-DC converter delivering V_O = 12V at I_O = 1A from a 24V input. The static input resistance is 24²/(12×1) = 48Ω.

One reason for having a regulator is to keep V_O constant as its source voltage varies. Given that $P_I = V_I I_I$ and P_I is constant, any increase in V_I will result in a proportional decrease in I_I. Under these conditions we need to consider the dynamic or incremental resistance. To do this, we assume that the input voltage changes by ΔV – that is, it changes from V_I to $(V_I + \Delta V)$. We can then use $I_I = V_O I_O/V_I$ to find the change in input current as it changes from $V_O I_O/V_I$ to $V_O I_O/(V_I + \Delta V)$. Using this in the definition of incremental resistance we get:

$$r_I = \frac{\Delta V}{\Delta I} = \frac{(V_I + \Delta V) - V_I}{\dfrac{V_O I_O}{(V_I + \Delta V)} - \dfrac{V_O I_O}{V_I}}$$

After some algebraic manipulation we get:

$$r_I = -\frac{(V_I + \Delta V) V_I}{V_O I_O}$$

ΔV is very small (tends to zero for dynamic resistance), so we can simplify this to:

$$r_I = -\frac{V_I^2}{V_O I_O} = -\frac{V_I}{I_I}$$

Which, by comparison with the equation above, can be seen to be the negative of the static resistance. This is a simplified analysis that assumes perfect regulation and 100% efficiency. In practice, the regulation is not perfect and the efficiency is not only less than 100%, but will vary with input voltage.

An important point here is the relationship between what the regulator is doing (in terms of regulation activity) and the effective input resistance of the regulator. To understand this we need to know something about how the regulator works. This will vary depending on the type of regulator, so we will just look at one example.

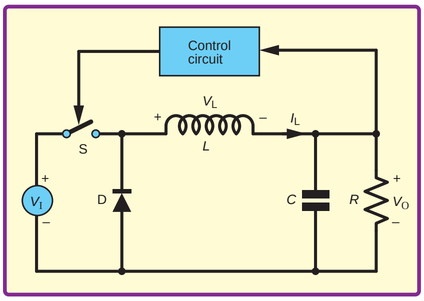


Fig.2. Basic switching converter schematic (buck converter)

Regulator operation

We will consider a switching regulator DC-DC converter that uses pulse-width modulation to control the output voltage. A simplified example circuit is shown in Fig.2 – this is a step-down buck converter ($V_I < V_O$). When the switch (S) closes, current flows from the source to the inductor and through the load. The inductor acts as an energy store (stored in its magnetic field). When the switch is open, the stored energy is released into the load, with the forward-biased diode providing the complete circuit for current to flow. When the switch is closed and the input voltage is connected across the diode, the diode is reverse biased and does not conduct.

When the switch closes the voltage across the inductor is $V_I - V_O$. If we assume V_O is constant (that is the job of the regulator) and V_I is also constant (for the situation under discussion) then the voltage across the inductor is constant at $V_I - V_O$. Applying a constant voltage to an inductor results in a constant rate of increase in current. This comes from the fundamental equation for the voltage-current relationship of an inductor:

$$V = L\frac{di}{dt}$$

Where i is the inductor current at a given time (t), V is the voltage across the inductor, L is the inductance. Again, we have differential equation: di/dt is the rate of change of current. If V and L are constant we can use $\Delta I/\Delta T$ instead of di/dt. From this we can write the inductor current change as:

$$\Delta I = \frac{V}{L}\Delta T$$

When the switch is open the voltage across the inductor is also constant and is equal to the load voltage, but opposite in sign (see the voltage orientations denoted on Fig.2). Under these conditions the inductor is outputting energy into the load and its current is ramping down at a constant rate – a negative slope of di/dt in the above equation corresponds with the negative voltage. Note that the inductor current flows in the same direction irrespective of switch position, but the relative polarity of the inductor and load voltage change by virtue of the switching arrangement. The switched nature of the circuit means that the output voltage has a tendency to vary (ripple) with the switching cycle, but this is smoothed by the capacitor (C).

The control circuit manipulates the time the switch is on in order to maintain the correct output voltage. The switch is activated on a cycle of constant duration T, so it is on for time DT, where D is the duty cycle – the fraction of time the switch is on for. In each cycle the switch is off for time $(1 - D)T$. The timing of the switch and corresponding inductor current waveform is shown in Fig.3. The control circuit varies the duration of the voltage pulse into the circuit, so this is a form of pulse-width modulation (PWM) control.

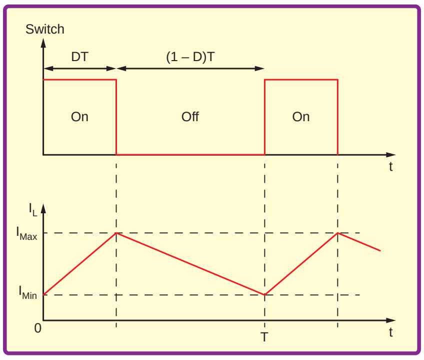


Fig.3. Timing and waveforms for the circuit in Fig.3

With the help of Fig.3 we can find the change of current $\Delta I = (I_{max} - I_{min})$ for the two parts of the cycle using the equation for ΔI given above. With the switch closed $\Delta t = DT$ and we have VI – V_O across the inductor, so

$$I_{max} - I_{min} = \frac{(V_I - V_O)}{L}DT$$

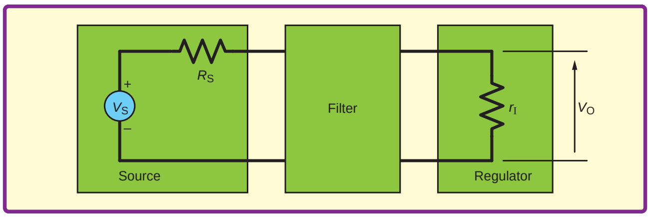


Fig.4. Regulator with input filter

With the switch open $\Delta t = (1 - D)T$ and we have $-V_O$ across the inductor, so:

$$I_{min} - I_{max} = \frac{-V_O}{L}(1 - D)T$$

As can be seen from Fig.4, the inductor current changes by the same amount in each part of the cycle. For steady operation we can equate the two ΔI equations:

$$\frac{(V_I - V_O)}{L}DT = -\frac{-V_O}{L}(1 - D)T$$

With a little algebraic manipualtion, this simplifies to:

$$\frac{V_O}{V_I} = D$$

Thus the control circuit will manipulate the duty cycle to set the output voltage. Given that (assuming 100% efficiency) $P_O = P_I = V_O V_I = I_O I_I$ so $V_O/V_I = I_I/I_O = D$ (using the equation above) we can also write $I_I = DI_O$ and $VI = V_O/D$. Returning to the equation for dynamic resistance and substituting for V_I and V_O we get:

$$r_I = -\frac{V_I}{I_I} = -\frac{V_O/D}{I_O D} = -\frac{1}{D^2}\frac{V_O}{I_O} = -\frac{R_L}{D^2}$$

Thus, the dynamic input resistance is determined by the regulator's duty cycle and the load resistance. Similar equations can be obtained for regulators operating on different principles.

Input filter and capacitors

A low-pass filter is commonly inserted between the source and regulator input (see Fig.4). The purpose of this filter is to prevent any high frequency noise and interference from reaching the regulator. The filter also prevents the regulator from causing electromagnetic interference (EMI), either by conduction via the circuit, or by radiation. It blocks high frequencies from the regulator to isolate its switching noise from other circuits and wiring which might radiate signals. The filter can be implemented by just a capacitor across the regulator inputs, or as a simple LC filter, as show in Fig.5, but a number of other variations are possible.

Fig.5 shows a single input capacitor, however switching regulators often have multiple capacitors on their input; reflecting multiple roles for the capacitor(s) to:

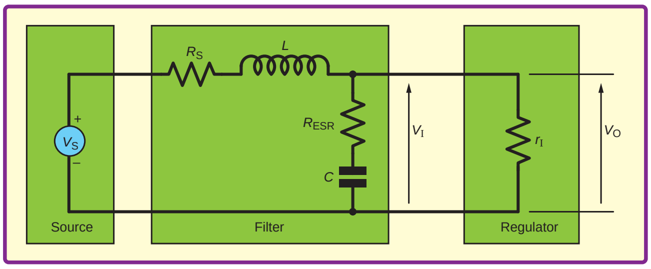


Fig.5. Fig.6. Regulator with example filter circuit model. This is a two-component LC filter, but the capacitor's ESR and the source resistance/inductor resistance will also influence the filter's characteristics.

- Smooth input voltage variation due to switching (ripple)
- Help cope with transient demands in load current
- Provide high-frequency bypass for the internal control circuits (for stability)
- Help shape the overall frequency response of the input filter.

Selection of capacitors is not simply a matter of the correct capacitor value. Real capacitors have effective series inductance and resistance (ESR), possible voltage dependence and ripple current ratings that need to be considered, along with physical and thermal characteristics. Thus, particular types of capacitor are likely to be more suitable than others and various trade-offs may need to be thought about. Typically, regulator datasheets provide guidance on capacitor selection. Multiple capacitors may be used because the imperfections and limitations of real components mean that one component cannot cover everything.

To look at ripple smoothing in a little more detail, we note that the regulator demands a varying input current from the source at the switching frequency (as shown in Fig.3), which in turn causes voltage variations due to inductance and resistance in the external circuit. These voltage and current variations are a potential source of EMI, as noted above. Connecting a suitable capacitor across the regulator input, as close as possible to the input, creates a bypass path at the relevant frequency, which shorts out any ripple. Typically, ceramic chip capacitors of units to tens of microfarads are used here, but when using a particular regulator IC, the datasheet should be consulted for details of capacitors selection.

When the regulator is responding to abrupt increases in load current it will in turn increase its input current demand. It is often the case that the source will not be able to respond quickly enough to meet this demand. Again, the effect of current variation is to cause voltage variation on the input, which in this situation is potentially large enough to disrupt regulator operation. The problem is most likely to occur if there is long wiring from the source to the regulator input and can be overcome by using a sufficiently large capacitor across the inputs, referred to as a bulk capacitor. Electrolytic capacitors of tens or hundreds of microfarads may be used for this purpose, but again, the datasheet may provide more specific guidance.

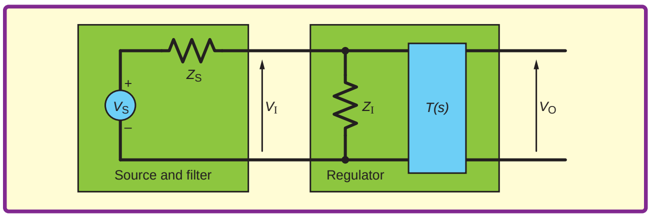


Fig.6. Regulator with example filter represented in the source impedance.

Source impedance and stability

The filter and source resistance can be combined into a Thévenin-style source impedance, as shown in Fig.6. We use Z_S rather than R_S to indicate that we dealing with a frequency-dependent impedance rather than resistance. Similarly, we represent the input to the regulator with impedance Z_I. The regulator's control circuit is represented by its transfer function $T(s)$, in which s is the Laplace complex frequency variable. As Martin indicates, a full analysis of the stability of the regulator's control circuit typically involves significant amount of mathematics, which is beyond the scope of this article.

We can get some insight into possible stability problems using Fig.6 without a full analysis. The overall relationship between the source voltage V_S and the regulator output is influenced by both $T(s)$ and the impedances Z_S and Z_I. In fact, Z_S and Z_I form a potential divider providing V_I to $T(s)$, so we can write:

$$\frac{V_O}{V_S} = \left(\frac{Z_I}{Z_S + Z_I}\right)T(s)$$

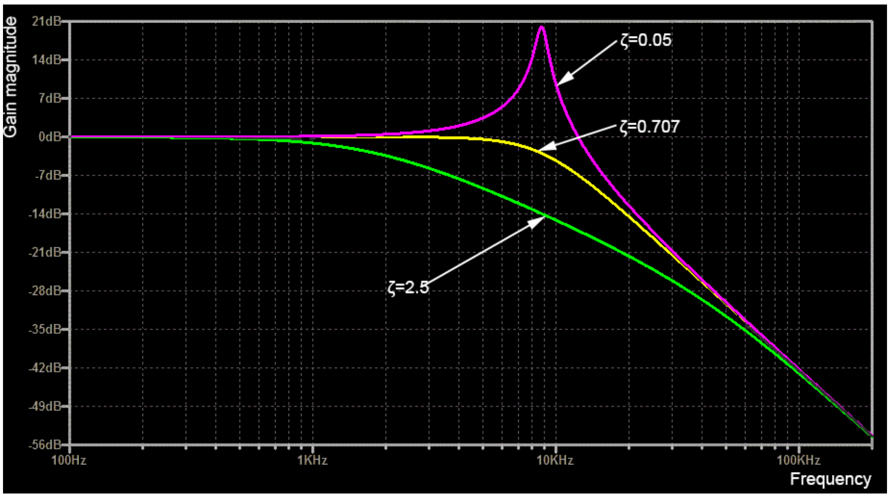


Fig.7. Typical filter response of an RLC second-order low-pass filter. This graph illustrates the general shape of the response and is not related to any specific regulator design

From this, we can conclude that if Z_S is much smaller than Z_I, then $Z_I/(Z_S + Z_I)$ will be close to unity and the relationship between V_O and V_S will be dominated by the control circuit $T(s)$. Under these conditions we assume the control circuit will be stable, but if Z_S becomes significant the behaviour of the system may change, with instability being a possibility. This leads to the rule of thumb that Martin mentioned that the magnitude of the source impedance should be at least ten times smaller than the magnitude of the regulator input impedance.

If the filter in Fig.5 is used, assuming $R_S = 0$ for simplicity, then we have a second-order RLC low-pass filter. The frequency response (see Fig.7) of this filter depends on the relative values of the components. Of particular important here is the parameter known as 'damping' (symbol ζ). If the damping is relatively high then the filter gain will decrease slowly past the cut-off frequency. With less damping the response will exhibit a peak near the cut-off frequency. This will effectively amplify the noise in the circuit at that frequency, and the condition discussed above ($Z_S << Z_I$) may not be met, so instability may occur. Suitable filter design can avoid this problem and there are a number of technical articles from semiconductor manufacturers that discuss regulator input filter design in detail.

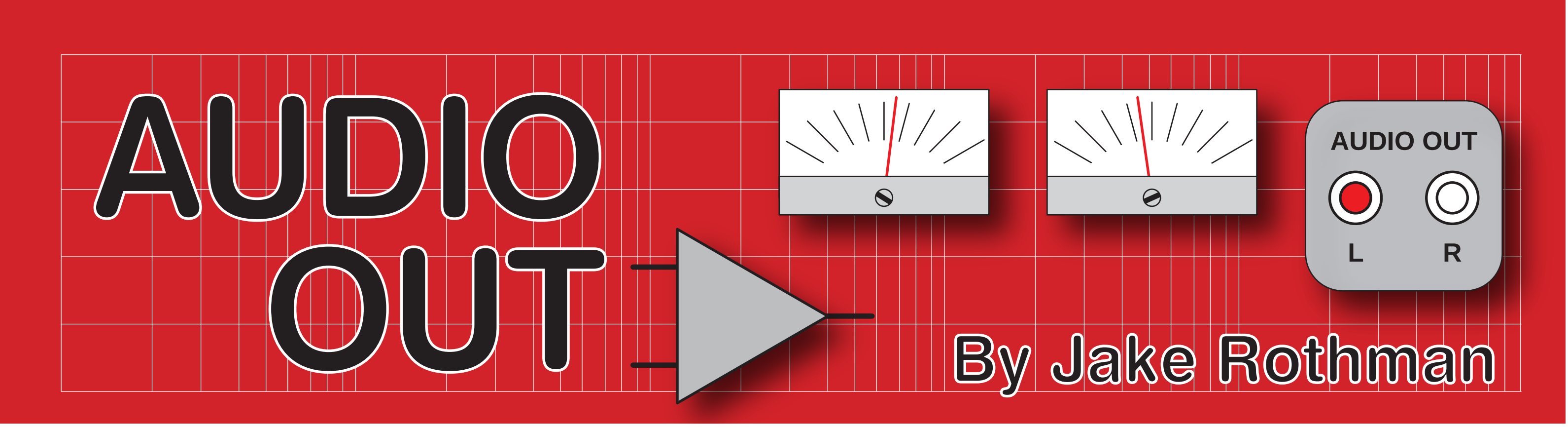
If one performs a stability analysis of the system in Fig.6, with the filter from Fig.5 (again with $R_S = 0$) conditions for the relationship between r_I and the filter components to ensure stability can obtained. The results (for the particular situation discussed) are:

$$|r_I| > R_{ESR}$$

$$|r_I| > \frac{L}{CR_{ESR}}$$

If the resistance (R_{ESR} in this case) is low then these conditions may not be met – also, from another perspective, the RLC circuit damping will be low and the previously discussed condition ($Z_S << Z_I$) may not be met at all frequencies.

Martin has found 'highly mathematical' treatments because this is a complex topic – so we cannot provide a simple comprehensive answer, except perhaps to say read the datasheet of any switching regulator you use carefully and follow any component selection and board layout guidelines provided.



Variable audio filters – Part 2: voltage-controlled filters

Last month, we looked at *manually* controlled variable corner frequency filter circuits that used potentiometers to change a filter's parameters. This month, we'll look at *voltage*-controlled filters (VCFs), which are the most important processor circuits in sound synthesisers. In VCFs a voltage rather than a change in a physical resistance controls the corner frequency.

VCFs use the same technology as voltage-controlled amplifiers (VCAs) which are normally used to control gain or volume. The simplest versions use an LED shining on light-dependent resistors, FETs and diodes, which effectively give voltage-controlled resistance. The problem with this technique is that there are large individual variations in resistance and control curve with such devices. The tracking between the elements is worse than that between potentiometer sections. Better, but more complex control devices use the variation of transconductance with current exhibited by bipolar transistors in fully balanced or differential configurations.

Breakthrough

Accurate balancing is important to prevent the control voltage itself from appearing on the output. This is called 'control voltage breakthrough' and can manifest itself as thumps when the control voltage is rapidly changed. Since the control voltage is normally in phase with the signal it has to be cancelled by a differential amplifier. Transistor matching of V_{be} and H_{fe} is important to ensure balance in these circuits. This is where chips and dual transistors have an advantage because the transistors are fabricated on the same piece of silicon and hence accurately share semiconductor characteristics.

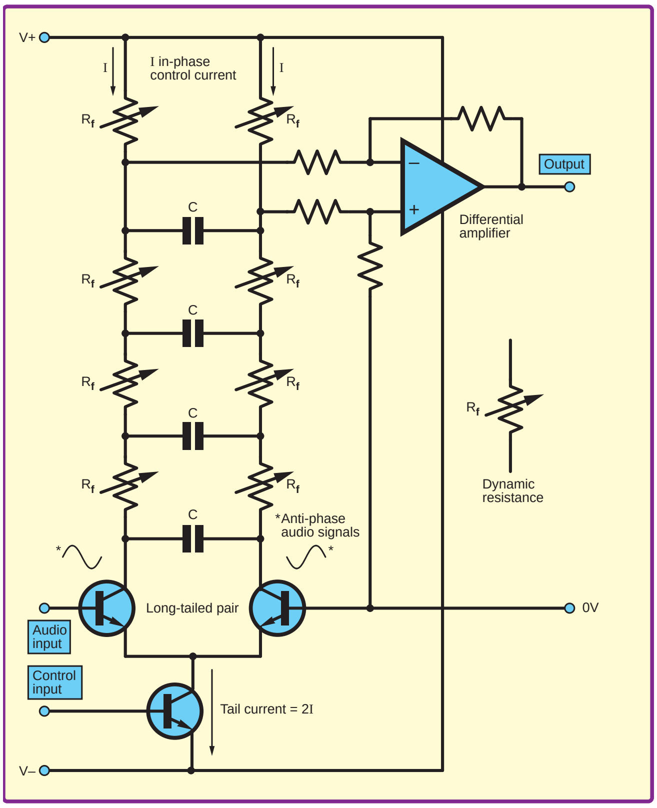


Fig.18. Structure of a ladder filter.

Bob Moog filter

The Moog synthesiser filter must be the most famous filter in music history. It has a unique circuit design and sound, which is possibly why it was the only part of Moog's Minimoog synthesiser that was patented (in 1966). (A little piece of *Audio Out* trivia – 'Moog' is pronounced 'mohg' with an 'oh' sound, not a bovine 'moo'. I discovered this from the horse's mouth when my friends and I were corrected by the late Robert Moog at a Theremin convention in 1996.)

Moog's circuit evolved from Blumlein's long-tailed pair differential amplifier. The two collector-load resistors are replaced with a resistor/capacitor ladder shown in Fig.18. The output is derived from a differential amplifier placed at the top of the ladder.

To vary the frequency, the value of all the ladder resistors has to be varied simultaneously. This is achieved using a circuit element to replace the resistors whose resistance varies with current. By modulating the current though the 'tail' (the bottom common-emitter resistor of the transistors) the current through both sides of the ladder is controlled, thus varying the cut-off frequency. The simplest component for this job is a diode, whose non-linear forward resistance is dependent on the current flow – see Fig.19. This resistance is the forward voltage (V_f) divided by the forward current I_d, often called 'dynamic resistance'. This is not very consistent between individual diodes, so instead, Bob Moog used individually biased transistors.

Control breakthrough is eliminated by cancellation in the differential output stage. The filtered signal is amplified rather than rejected because the audio outputs of the ladder are out of phase.

In music synthesis fourth-order filters are normally used to get a more dramatic effect, which means four capacitors are needed, along with eight transistors. Fig.20 shows a Moog ladder filter and a prototype built on Veroboard (Fig.21). Note the use of a CA3046 transistor array to ensure the lower and upper long-tailed pair transistors are matched.

The Gen X-1 I recently designed (Fig.22a) uses the cheaper relation of the Moog ladder filter, the diode ladder filter. The famous 1974 *Practical Electronics* Minisonic and Gakken SX-150 synthesisers also use this circuit. In the Gen X-1 filter, the diodes are transistors wired as diodes, which exhibit a more consistent forward voltage. The top of the diode ladder must be isolated from power supply

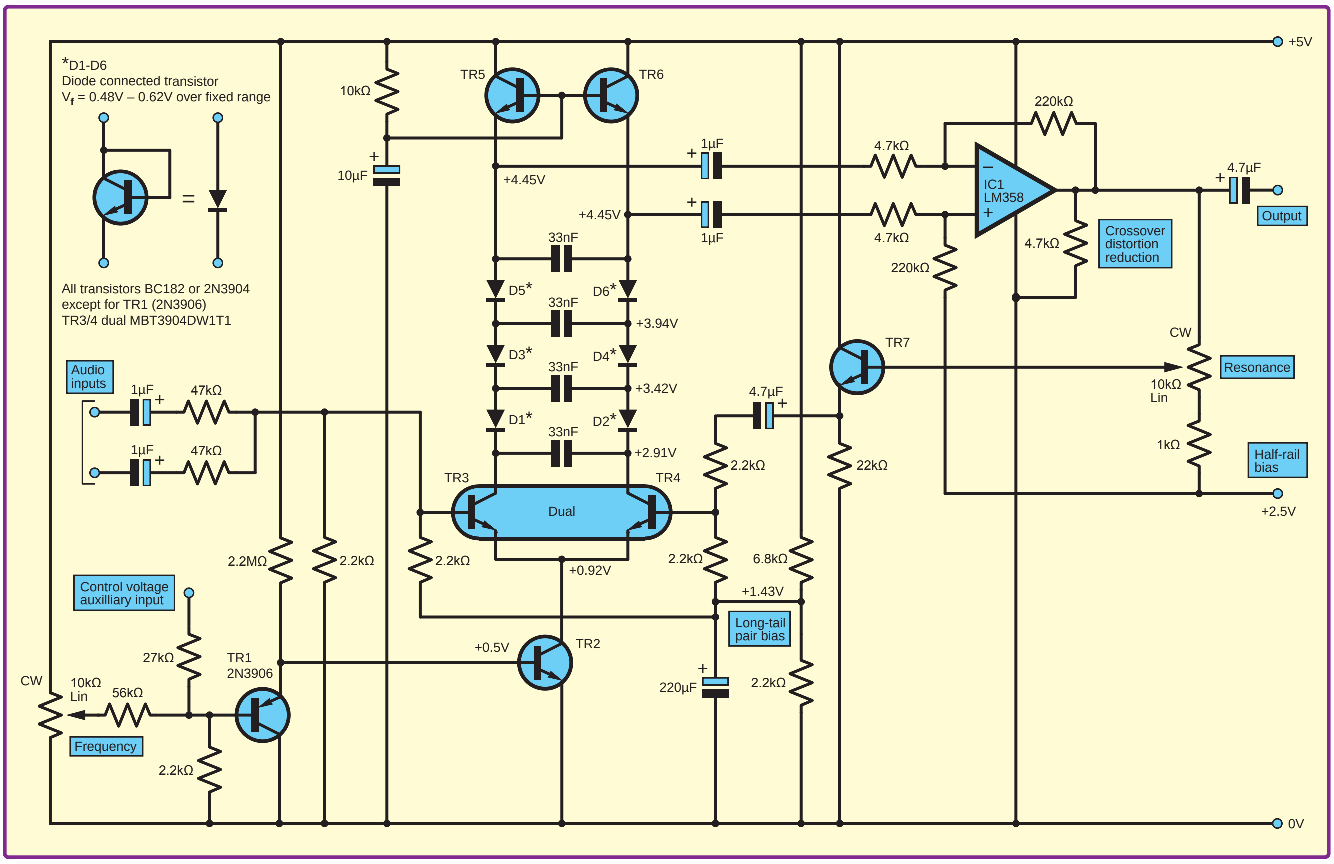


Fig.19. Diode ladder filter used in the Gen X-1. This actually uses transistors wired as diodes, since the matching of the forward-biased resistance characteristic is better.

noise, hence the use of a pair of emitter followers built around TR5 and 6 with decoupled biasing. The construction is surface-mount, where the devices cost pennies, and is shown Fig.22b. The lower long-tailed-pair transistor is a dual version of the 2N3904, which costs around 10p and ensures the ladder currents are equal. The moral of this story is that it's now cheaper to build analogue circuitry with discrete surface mount technology than using single-sourced chips.

Positive feedback

For musical synthesis, it is useful to be able to put plenty of positive feedback around the filter to make it highly resonant, often to the point of oscillation. This gives synths their characteristic 'wah' sound as the filter opens up. Ladder filter circuits do this very well because the gain is reduced proportionally as the Q increases. The circuits also soft-clip and there is intermodulation between the input signal and the filter resonant frequency in a unique manner that is recognised as the 'Moog sound'.

Transconductance op amp filters

There are a lot of components in a ladder filter, which makes them difficult to assemble by hand, especially on stripboard. The solution is to use an IC, which will reduce the parts count considerably. The most popular devices for filters are operational transconductance amplifiers or OTAs. They are basically op amps with a gain control pin. The output is a current output (hence the current source in the symbol) as is the gain control, which is a current input. The output current is the product of the input voltage times the control current, thus giving gain control. OTAs are normally used open loop; that is, with no negative feedback. If negative feedback were used it would act to even out the desired gain variation. The resulting high gain of open-loop operation means only about 50mV peak-to-peak can be applied to the input before clipping occurs. This means an attenuator has to be inserted on the input.

The most popular OTA chip in the heyday of the analogue synth, the 1980s, was the CA3080 and the famous Oberheim synthesiser used two of them for its filter. Unfortunately, transconductance amplifiers such as the CA3080, CA3060 and the higher quality CA3082 are now expensive because they are no longer made and eBay sellers are taking advantage of demand and scarcity. The chips used the now obsolete 7µm-fabrication process. It seems there will be no new plant commissioned, a disaster for synth builders. However, for the moment, you can buy the LM13600/700, which is still available from National Semiconductor and JRC. This chip consists of a couple of 3080 masks combined with a pair of Darlington transistors available for use as buffers. This enables a whole filter to be built with one chip. There are also linearising diodes on the inputs that reduce the distortion. These work by generating a distortion opposite to that produced by the chip's input transistors. This cancellation only works up to a certain point, at which point the distortion suddenly returns.

An LM13600 OTA filter circuit is given in Fig.23. It is basically a form of state variable with band-pass and low-pass outputs. The output current can feed directly into a capacitor, forming an integrator. A buffer stage is then required after the capacitor. This block can then be put into the state-variable filter topology. With the CA series chips, the buffer can be an op amp wired as a voltage follower or a JFET source follower. The Darlington buffers in the LM13600 have their operating current linked to the control pin current to maintain a higher input impedance. In theory, this should be an advantage for filters since the drain on the integrating capacitor should be less at low control currents. The LM13700 chip does

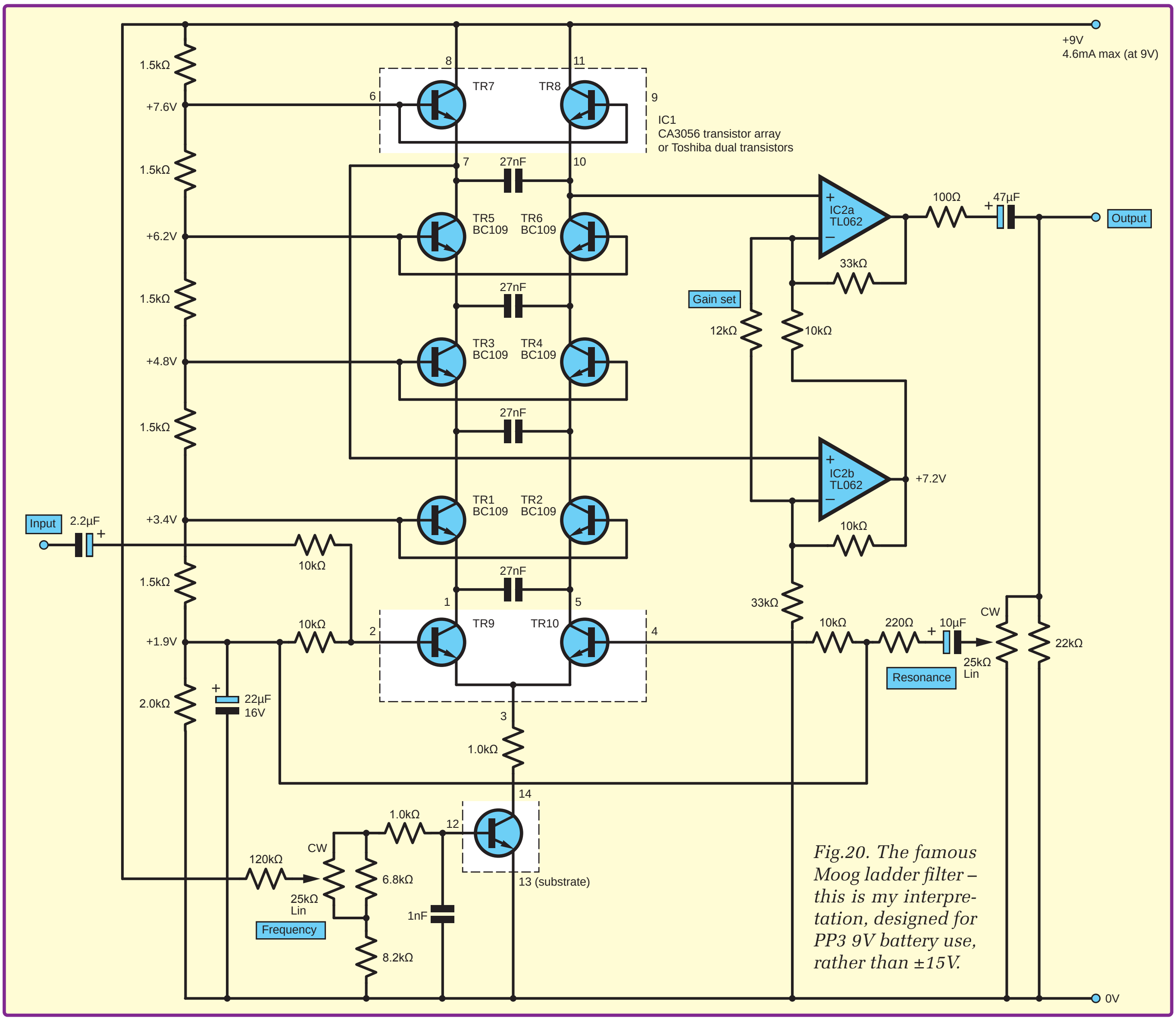


Fig.20. The famous Moog ladder filter – this is my interpretation, designed for PP3 9V battery use, rather than ±15V.

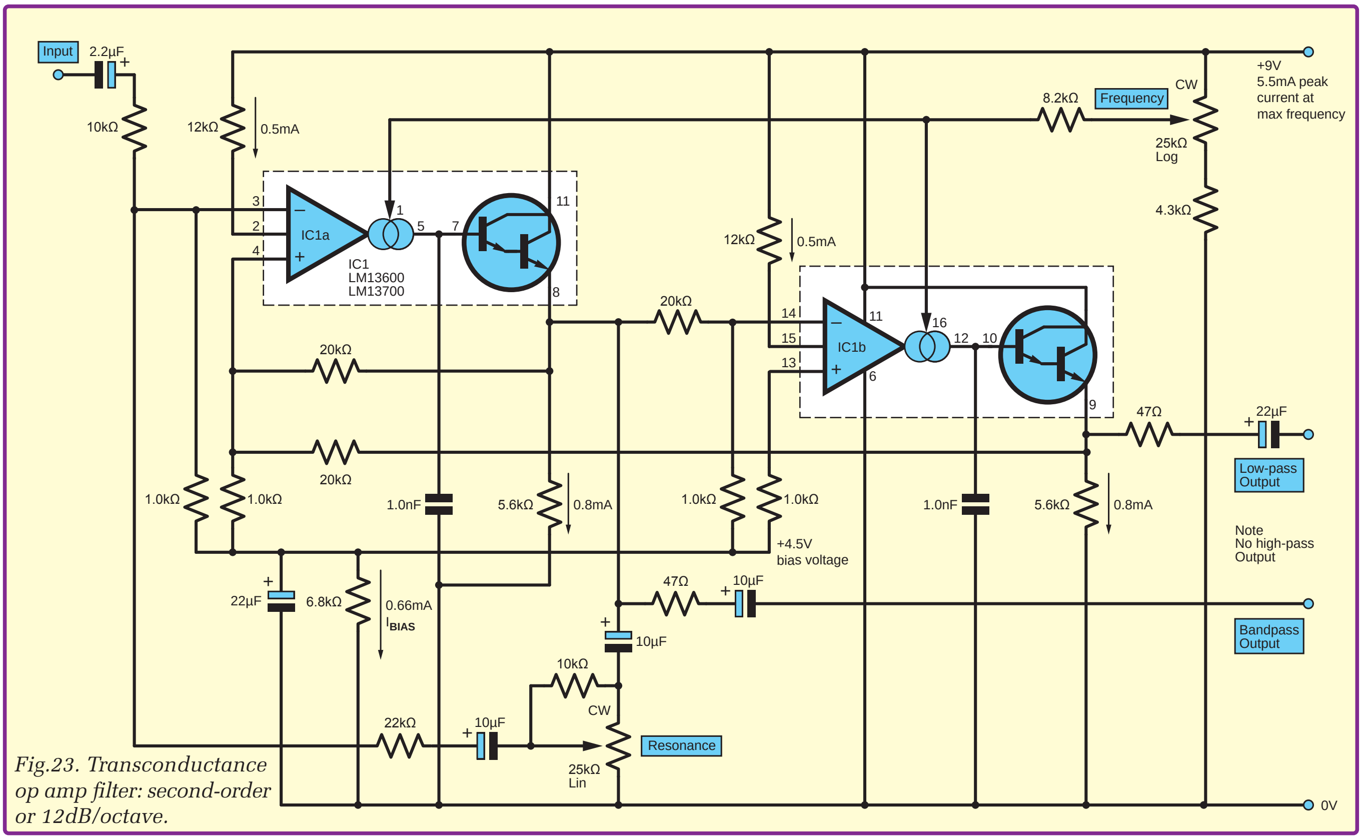


Fig.23. Transconductance op amp filter: second-order or 12dB/octave.

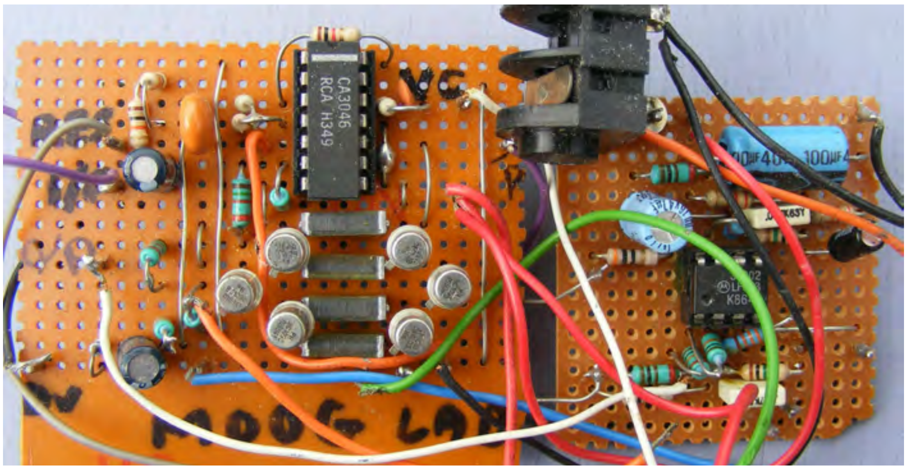


Fig.21. The Moog ladder filter built on Veroboard.

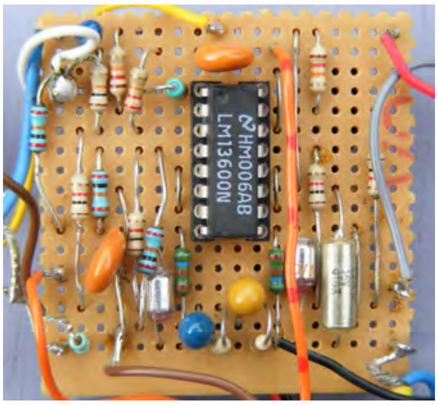


Fig.24. Veroboard construction of the transconductance op amp filter in Fig.23.

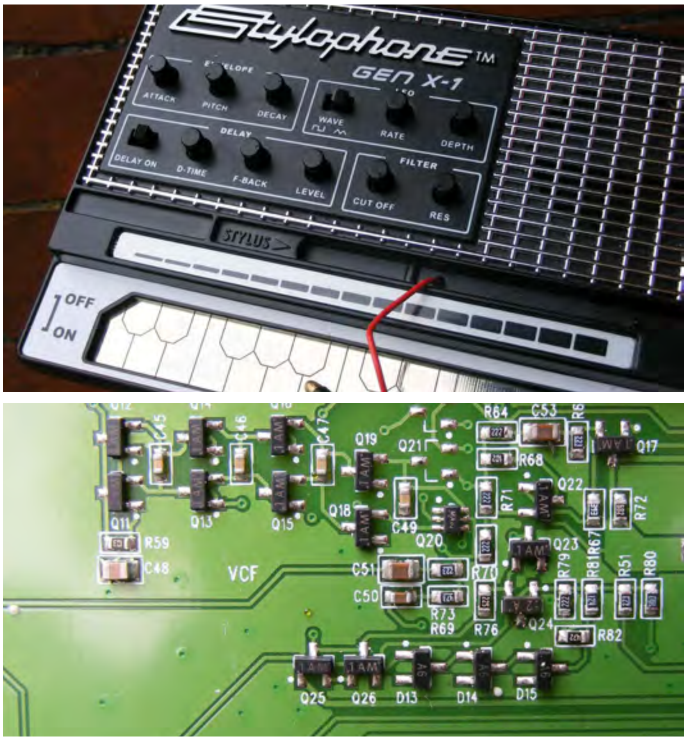


Fig.22. a) Gen X-1 synthesiser; b) Inside the synthesiser: the diode ladder filter made from discrete transistors can be seen. The total cost of the entire filter is less than an LM13600 and can run on 5V.

not have this feature, but I noticed no difference when used with this filter. The chip has the advantage of being available in surface mount.

Unusually, this circuit runs on a PP3 9V battery – the standard choice for guitarists and students, who find lugging mains dual-rail PSUs around a pain. In this circuit the current from the linearising diodes also generates the centre bias voltage across R20. The chips are quite noisy and not really 'Hi-Fi', but are effective in synthesisers or as a musical effect. Although the distortion is high at around 0.6%, it does not sound objectionable. It is more a form of soft clip. Although this design is only second-order, it is truly so, giving a 12dB/octave roll-off (the ladder filters giving a few dB less than their theoretical 24dB/octave).

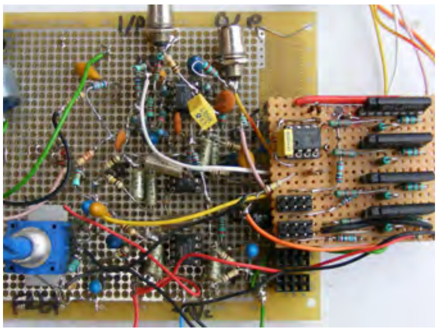


Fig.27. A prototype of the fourth-order filter. Note the plug-in board with four SIL (single in line) VCAs. The type used were the dbx 2180 (now made by That Corporation) available from Profusion in the UK. This was designed to evaluate the performance of different control elements.

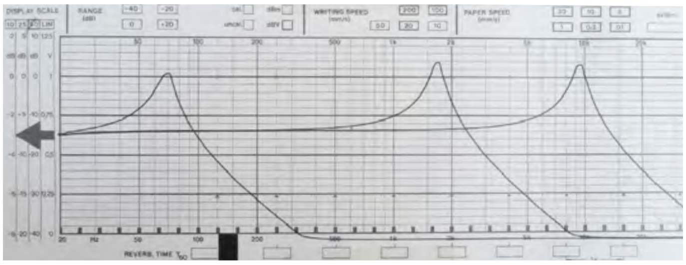


Fig.25. Peaky synthesiser response obtained by boosting the resonance.

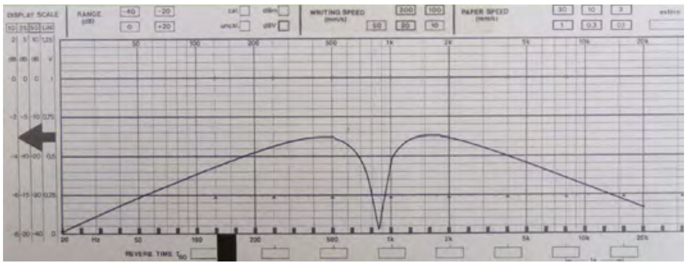


Fig.28. The notched band-pass filter output.

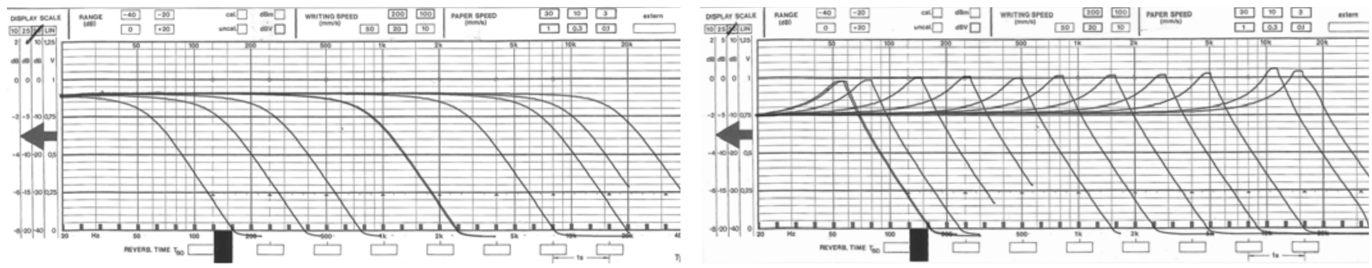


Fig.29. a) Low-pass response of the Blackmer VCA filter. Note consistency of curve as control voltage is varied. This would not be possible with a normal multi-gang potentiometer. Since this is a fourth-order filter the scale has been changed to 50dB or 2dB per a division; b) VCA filter at max Q. The slight droop in peak height towards the bass is due to the AC coupling

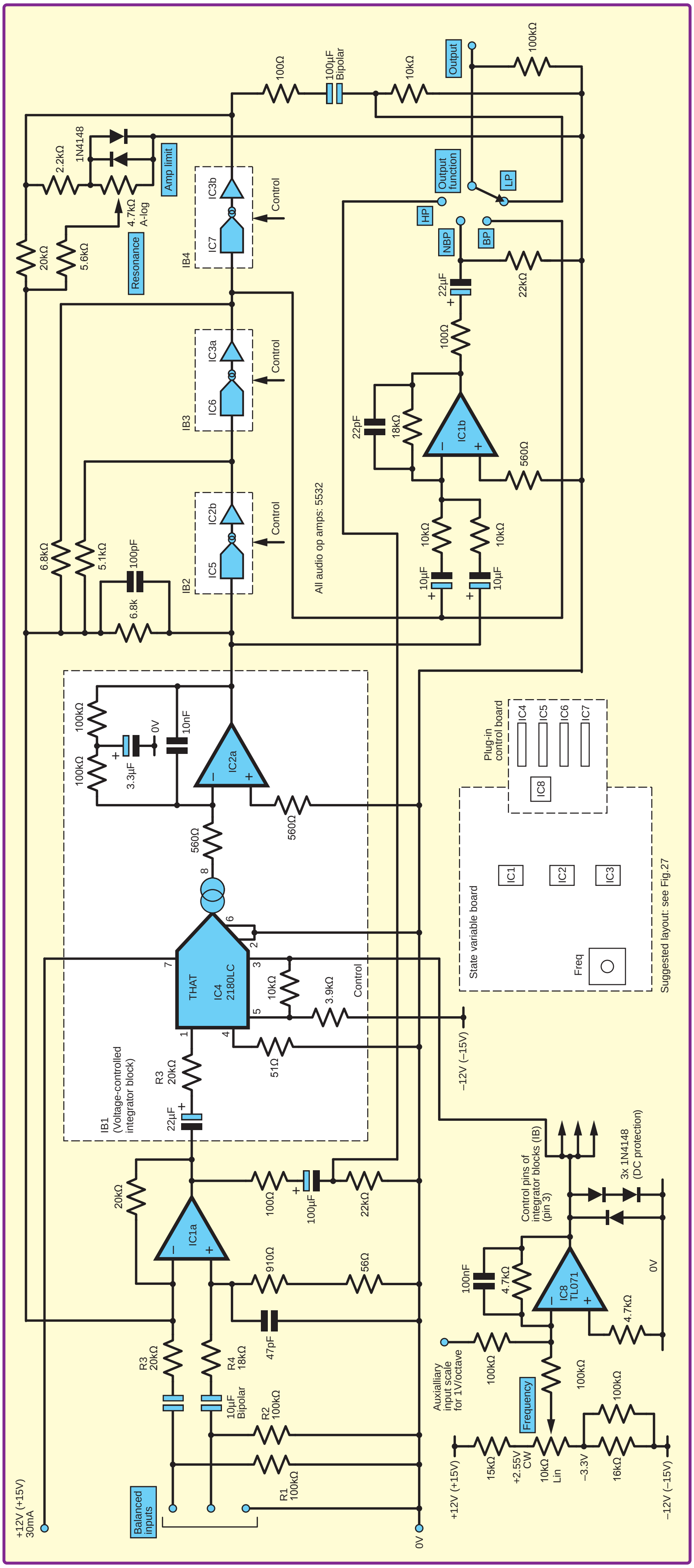


Fig.26. Fourth-order filter using 2180 Blackmer VCA chips from That Corporation as the control elements. Complex, but capable of full Hi-Fi performance. It avoids having to obtain an 8-gang precision anti-logarithmic potentiometer for stereo use.

A linear potentiometer loaded by a low-value resistor provides a logarithmic frequency control.

A photo of the simpler Veroboard construction is shown in Fig.24. Fig.25 shows the effect of turning up the resonance. It is possible to cascade two LM13600 filters to get a fourth-order filter, but the resonance control becomes difficult because of the higher phase shift. A dual-gang potentiometer could possibly be used with one for each second-order section, but this will impair reliability. In the fourth-order Oscar synthesiser filter, the resonance function was made voltage-controlled by adding an additional LM13600, allowing use of a single-gang potentiometer. I plan to try making a third-order filter with the spare section for a VCA. This would simplify the construction of a small synthesiser for *EPE* one day.

Mega-posh filter

My favourite filter is shown in Fig.26; it has proper Hi-Fi performance. It is based on the fourth-order state-variable configuration shown last month, but with Blackmer VCAs for frequency control rather than a potentiometer. As a result, the tracking of the integrator sections is much better, resulting in a more consistent response curve. A single potentiometer can be used to control all eight integrator sections in a stereo filter. The cumulative distortion of the VCAs is very low, in the order of 0.15% and subjectively benign. A full orchestra can be put through it and still sound clean and smooth. Blackmer VCAs are inverting, so the feedback paths from the first and third integrator stages have to be fed into the inverting pin of the summation stage IC1a, unlike the original fourth-order state variable. If this isn't observed the result will be an oscillator. The prototype circuit is shown in Fig.27. It has an additional output which is the two band-pass outputs summed together which results in a band-pass with a notch in the middle shown in Fig.28. This provides a unique phaser type sound when the frequency is swept up and down. Fig.29a shows the curve across the audio range and Fig.29b shows a degree of peaking applied.

When I get a PCB done, I'll do this circuit as a proper project. DJs and live electronic musicians will love it. It makes a fantastic filter swell pedal.

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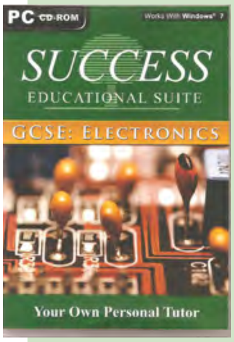

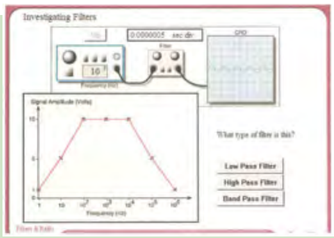

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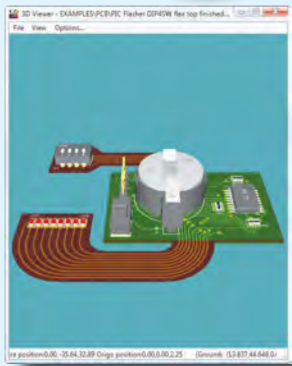

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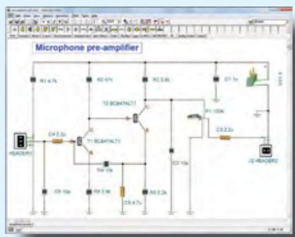

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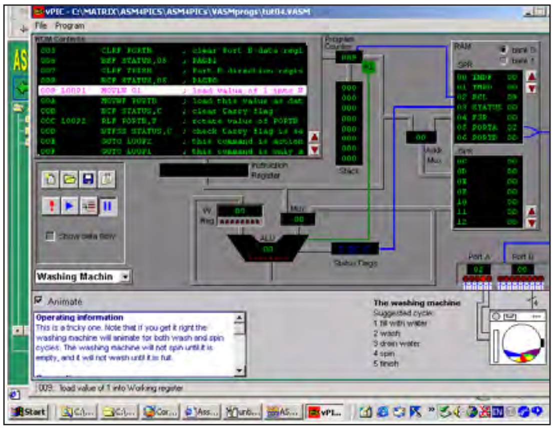

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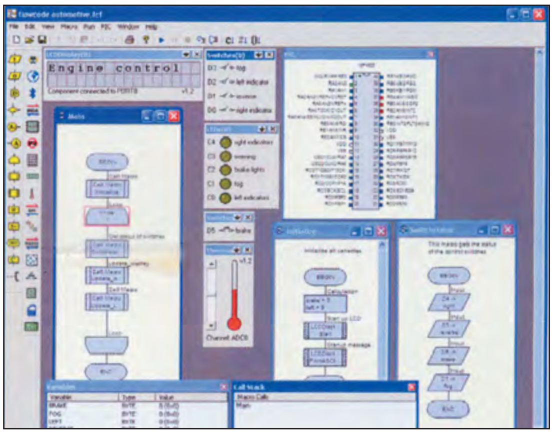

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This *Teach-In* series of articles was originally published in *EPE* in 2008 and, following demand from readers, has now been collected together in the *Electronics Teach-In 2* CD-ROM.

The series is aimed at those using PIC microcontrollers for the first time. Each part of the series includes breadboard layouts to aid understanding and a simple programmer project is provided.

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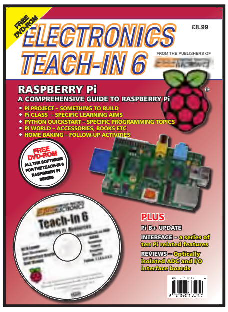

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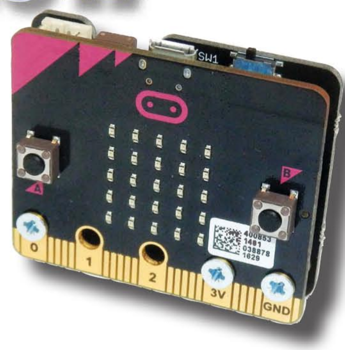

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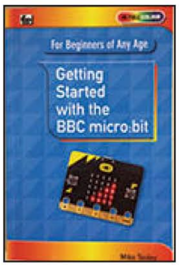

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Mike Tooley

Not just an educational resource for teaching youngsters coding, the BBC micro:bit is a tiny low cost, low-profile ARM-based single-board computer. The board measures 43mm × 52mm but despite its diminutive footprint it has all the features of a fully fledged microcontroller together with a simple LED matrix display, two buttons, an accelerometer and a magnetometer.

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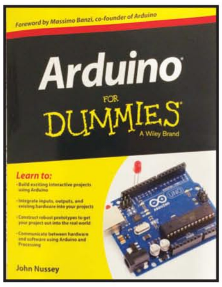
Teach-In 2016

Exploring the Arduino



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John Nussey

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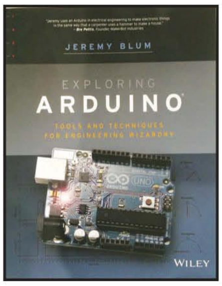

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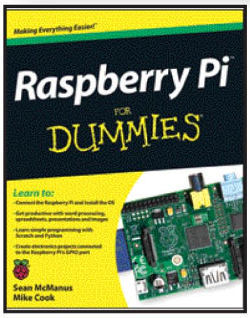

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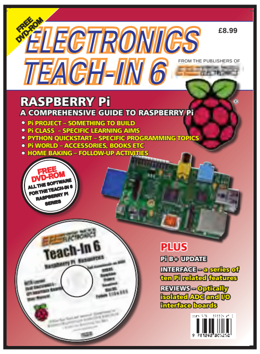

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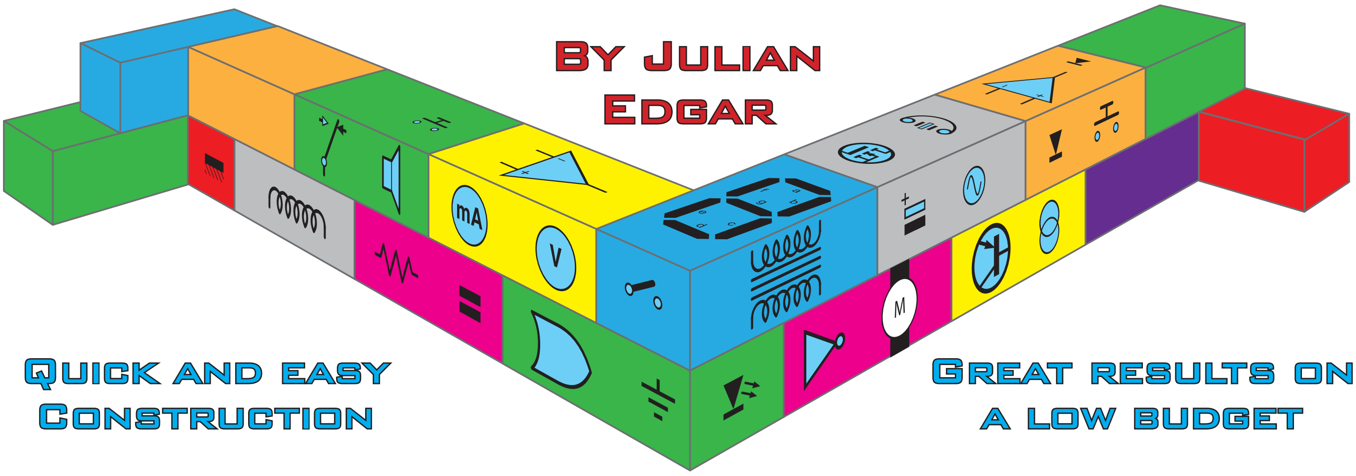
Electronic Building Blocks



By Julian Edgar

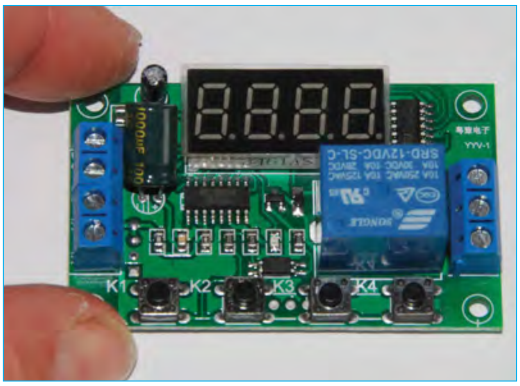
Quick and easy construction

Great results on a low budget

Digital Voltage Switch

Large complex projects are fun, but they take time and can be expensive. Sometimes you just want a quick result at low cost. That's where this series of *Electronic Building Blocks* fits in. We use 'cheap as chips' components bought online to get you where you want to be... FAST! They represent the best value we can find in today's electronics marketplace!

Has there ever been a better time to be pursuing electronics as a hobby? Take this module. Its basic function is a voltage switch – it activates the on-board relay when the DC voltage being monitored is higher (or lower) than the pre-set value. But it's more than that. It also has an LED display to show you the voltage being monitored, and seven different selectable switching modes, including the ability to precisely set the hysteresis (the difference between the 'on' and 'off' values). You can even switch off the LED display to save battery juice. It's versatile, accurate, cheap and effective!



This module can monitor DC voltages up to 50V, activating the on-board relay when the voltage reaches the required level. The module has seven different selectable modes, an LED display that shows monitored voltage, and adjustable hysteresis.

Note that while there are a lot of digital voltage switches available, they vary substantially in features. This particular one, identified by YYV-1 on the board, may not be the first that you come across when searching, so ensure the module you buy looks exactly like the one shown here. It should cost you around £5, including delivery.

But before we continue, of what use is a voltage switch? A simple application is as a low-voltage cut-out with battery-powered equipment. For example, if you like camping with a battery-powered light (eg, running off the car battery), this switch can be used to ensure that you don't unduly flatten the battery and so find it hard to start the car in the morning. The on-board relay is rated at 10A at 30VDC, and so you won't even need to add another relay. Another 'battery' use is to switch off an older, manual charger when battery voltage rises sufficiently.

You can also use the switch with suitable sensors to monitor temperature, pressure, light intensity, rotational position – you name it! Many such sensors directly output a voltage (eg, 1-5V) and can be connected directly to the voltage switch. Other sensors vary in resistance with the parameter being sensed (eg, temperature or light intensity) and so can be connected in a voltage divider configuration and fed via a regulated voltage. In some applications (eg, cars) there are numerous standard sensors that output voltages and so can be monitored, for example to sound alarms for over-temperature, low fuel, low oil pressure and so on.

The name of this column is *Electronic Building Blocks* – and this voltage switch is one of the best blocks to have on your shelf!

Connections

The module's connections are straightforward. Power and ground (12V in the module I used, but it's also available in 5V and 24V versions), and signal input and signal ground. (In this case, the signal is the voltage that we're monitoring.) Voltages up to 50V can be monitored. The relay is an SPST design, so it has common, normally-open and normally-closed connections. All connections are via screw terminals.

The measured power consumption of the module is as follows:

- LED display off – 16mA
- LED display on – 24mA
- LED display on and relay tripped – 56mA

NB: signal input impedance is ~20kΩ.

Modes

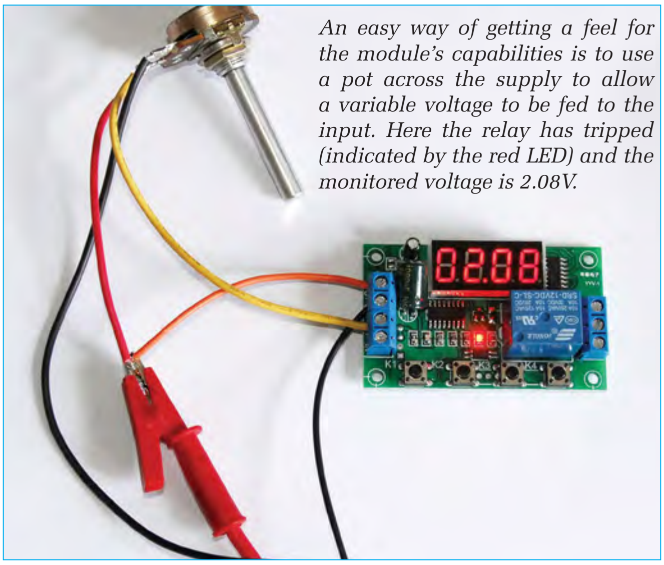
The module has seven mode configurations. The manufacturer's descriptions of these modes are only available in poorly translated English – so here they are in straightforward language!

Mode 1

When the input voltage is below V1, the relay is activated. When the input voltage is higher than V2, the relay is de-activated. This mode is typically used when you wish to monitor a falling voltage and want the relay to trip when the voltage drops below the set-point (V1). The difference between V1 and V2 sets the hysteresis.

Mode 2

When the input voltage is higher than V2, the relay is activated. When the input voltage is lower than V1, the relay is deactivated. This mode is typically



An easy way of getting a feel for the module's capabilities is to use a pot across the supply to allow a variable voltage to be fed to the input. Here the relay has tripped (indicated by the red LED) and the monitored voltage is 2.08V.

used when you are monitoring a rising voltage and want the relay to trip when the voltage rises above the set-point (V2). The difference between V2 and V1 sets the hysteresis.

Mode 3

In this mode the relay is activated only when the monitored voltage is outside the range set by V1 and V2. For example, if V1 is set to 6V and V2 is set to 7V, the relay is off when the measured voltage is between 6V and 7V, and is on when the voltage is less than 6V or greater than 7V.

Mode 4

This mode reverses the logic of Mode 3 – the relay is activated only when the monitored voltage is between V1 and V2.

Mode 5

This mode activates the relay when the monitored voltage is above V1 and switches off the relay when the monitored voltage is below V1. The difference between this mode and Mode 1 is that the hysteresis cannot be adjusted. In use, the hysteresis in this mode appears to be 0.05V. (V2 therefore does nothing, even though it still appears as an item to be set.)

Mode 6

When the monitored voltage exceeds V1, the relay is activated. The relay stays activated until power is lost to the module. This mode is particularly good when you need to monitor a signal and see if a parameter has been exceeded, for example, in your absence. A normally closed 'reset' button in the module's power supply could be easily added to switch off the alarm. Note that the LED display keeps showing the actual 'live' value, even if the relay has been tripped.

Mode 7

The relay is activated only when the monitored voltage is precisely at the set-point (V1). Even an indicated 0.05V variation away from V1 switches off the relay.

Mode selection

There are four on-board press-buttons, marked as K1 to K4. K2 is 'up' and K3 is 'down'. The mode is selected by holding down K1 for more than one second. The LED display will then flash – eg, P--2, indicating that Mode 2 is selected. To change the mode, press either the up or down buttons until the required mode is selected. Pressing K1 again will then show the V1 that is selected; this can be changed with the up and down buttons. Pressing K1 again will select V2, that can then be set. A final push of K1 returns the display to normal. And K4? This button turns the LED display on and off.

In use

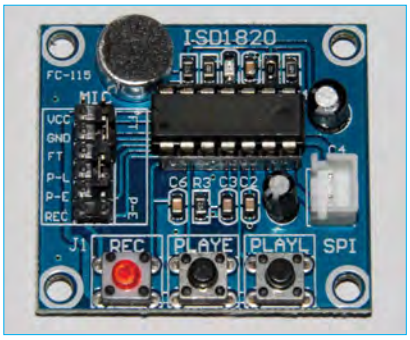
The set-point voltages can be with an apparent resolution of no less than 0.01V; however, this needs to be taken with a grain of salt. Varying the input voltage shows that the monitored voltage varies by a smallest increment of about 0.05V – so setting V1 or V2 to a smaller increment than 0.05V is ineffective. That said, being able to set the trip-point to 0.05V is still pretty good!

Note that the voltage regulation of the module is excellent: the measured input voltage holding dead steady at module supply voltages from 10-14V.

In use, it's usually not critical but for interest's sake, the displayed voltage also compared very well with that measured by a recently calibrated multimeter.

At the time of writing, the voltage switch is available on **www.aliexpress.com** (search under YYV-1) and from **eBay.co.uk** under the search term 'Charging Discharge Voltage Monitor Test Relay Switch Control Module DC 12V G9Q3' – currently, item 152846808873 for £3.96 incl p&p.

I like the modules so much that I bought six!



Next time

In my next column I'll be looking at a: *Pre-recorded sound module*.

PCB SERVICE

CHECK US OUT ON THE WEB

Basic printed circuit boards for most recent *EPE* constructional projects are available from the *PCB Service*, see list. These are fabricated in glass fibre, and are drilled and roller tinned, but all holes are a standard size. They are not silk-screened, nor do they have solder resist. Double-sided boards are **NOT plated through hole** and will require 'vias' and some components soldering to both sides. **NOTE: PCBs from the July 2013 issue with eight digit codes** have silk screen overlays and, where applicable, are double-sided, plated through-hole, with solder masks, they are similar to the photos in the relevent project articles.

All prices include VAT and postage and packing. Add £2 per board for airmail outside of Europe. Remittances should be sent to **The PCB Service, Everyday Practical Electronics, Wimborne Publishing Ltd., 113 Lynwood Drive, Merley, Wimborne, Dorset BH21 1UU. Tel: 01202 880299; Fax 01202 843233; Email: orders@epemag.wimborne.co.uk. On-line Shop: www.epemag.com**. Cheques should be crossed and made payable to *Everyday Practical Electronics* **(Payment in £ sterling only)**.

NOTE: While 95% of our boards are held in stock and are dispatched within seven days of receipt of order, please allow a maximum of 28 days for delivery – overseas readers allow extra if ordered by surface mail.

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*** See NOTE left regarding PCBs with eight digit codes ***

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A large number of older boards are listed on, and can be ordered from, our website.
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EPTSOFT Ltd 47
ESR ELECTRONIC COMPONENTS 44
HAMMOND ELECTRONICS Ltd 9
JPG ELECTRONICS 72
LASER BUSINESS SYSTEMS 51
MICROCHIP . . . . . Cover (ii), Cover (iii) & 6
PEAK ELECTRONIC DESIGN . . . . . Cover (iv)
PICO TECHNOLOGY 44
POLABS D.O.O. 55
QUASAR ELECTRONICS 2/3
STEWART OF READING 33
TAG-CONNECT 47

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Next Month

JUNE '18 ISSUE ON SALE 3 MAY 2018

10-Octave Stereo Graphic Equaliser

This stereo graphic equaliser is very compact and quite cheap to build. However, it has the performance to match full-blown commercial models that are far more expensive. It can also be used in a wide range of applications from AC or DC supplies.

Arduino-based Digital Inductance/Capacitance Meter

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Low Cost Electronic Modules – Part 5

Fancy using a 16x2 backlit alphanumeric LCD module? This one comes with a small 'piggy-back' module that provides it with an I^2C serial interface. It can be hooked up to any of the common micros via only two wires, letting multiple displays (or other I^2C devices) share the same 2-wire bus, while also freeing up some of the micro's I/O pins for other purposes.

Teach-In 2018 – Part 9

Next month's *Teach-In 2018* will bring the series to a close with some advice on designing and building your own test gear.

PLUS!

All your favourite regular columns from *Audio Out* and *Circuit Surgery* to *Electronic Building Blocks*, *PIC n' Mix* and *Net Work*.

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